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ELECTROTEHNICĂ
ENERGETICĂ
ELECTRONICĂ

1991

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ИЗВЕСТИЯ ЯССКОГО ПОЛИТЕХНИЧЕСКОГО ИНСТИТУТА
Tomul XXXVII (XLI), Fasc. 1—4 1991

Secţia III

ELECTROTEHNICĂ. ENERGETICĂ. ELECTRONICĂ

SUMAR

Pag.

D. I. MANGERON I

H. ROSMAN, Asupra realizabilităţii ferorezonanţei în circuite neliniare inerţiale, în regim permanent armonic (franc., rez. rom.) 1

S. MITREA şi A. ADĂSCĂLIŢEI, Asupra determinării capacităţii pentru evitarea ferorezonanţei serie (engl., rez. rom.) 5

I. CASIAN, Un mod de tratare a sistemelor cu constante distribuite (engl., rez. rom.) 11

I. SUCHAR şi R. URSULEAN, Asupra unui releu de proximitate fără contacte (engl., rez. rom.) 15

M. ANTONIU, E. VREMERĂ, M. BUZDUGA şi I. CHIABURU, Aparat pentru măsurarea electrizării triboelectrice a pulberilor izolante (engl., rez. rom.) 21

E. CREŢU, M. LUCANU, A. CREŢU şi C. FILIPIUC, Divizor inductiv de curent compensat electronic (engl., rez. rom.) 27

M. CREŢU, C. SĂRMĂŞANU şi L. BRENIUC, Asupra determinării poziţiei căruciorului mobil al macaralelor cu braţ orizontal şi afişarea sinoptică a rezultatului (franc., rez. rom.) 35

M. ADAM şi A. BARABOI, Controlul parametrilor dinamici ai echipamentelor electrice de comutaţie (engl., rez. rom.) 41

E. HNATIUC, Cu privire la proiectarea optimă a electromagneţilor de curent alternativ de tip U-I prevăzuţi cu spire în scurtcircuit (franc., rez. rom.) 45

V. MUNTEANU, M. LUCANU, A. PĂSTRĂVANU şi D. TĂRNICERU, Noi metode de decodare a codurilor binare liniare (I) (engl., rez. rom.) . . . 49

P. COTAE şi N. D. ALEXANDRU, Lăţimea de bandă ocupată de primul nul spectral al semnalelor de tip MSK utilizate în sistemele QAM (engl., rez. rom.) 57

M. S. OPRIŞOR şi S. A. CABA, Modelul matematic al frînării subsincrone cu condensatoare, a motorului asincron trifazat (engl., rez. rom.) 63

J. IGNAT, Metodă de identificare şi localizare a unei puneri la pămînt într-o reţea cu neutrul izolat (franc., rez. rom.) 71

GH. CĂRŢINĂ, Asupra evoluţiei consumului de energie electrică (franc., rez. rom.) 77

RECENZII 83

BULETINUL INSTITUTULUI POLITEHNIC DIN IAŞI
BULLETIN OF THE POLYTECHNIC INSTITUTE OF JASSY
ИЗВЕСТИЯ ЯССКОГО ПОЛИТЕХНИЧЕСКОГО ИНСТИТУТА
Tome XXXVII (XLI), Fasc. 1–4 1991

Section III

ELECTROTECHNICS. ENERGETICS. ELECTRONICS.

CONTENTS

	Pp.
D. I. MANGERON	I
H. ROSMAN, Sur la réalisabilité de la ferrorésonance dans les circuits nonlinéaires inertiaux en régime permanent harmonique (French, Romanian summary)	1
S. MITREA and A. ADĂSCĂLIŢEI, Upon the Determination of Capacitance to Prevent Series Ferroresonance (English, Romanian summary)	5
I. CASIAN, A Representation of Finite Distributed Systems (English, Romanian summary)	11
I. SUCHAR and R. URSULEAN, On Proximity Relays without Contacts (English, Romanian summary)	15
M. ANTONIU, E. VREMERĂ, M. BUZDUGA and L. CHIABURU, Apparatus for Measuring Triboelectric Electrification of Insulating Powders (English, Romanian summary)	21
E. CREŢU, M. LUCANU, A. CREŢU and C. FILIPIUC, Electronically Compensated Inductive Current Divider (English, Romanian summary)	27
M. CREŢU, C. SĂRMĂŞANU et L. BRENIUC, Sur la détermination de la position du chariot mobile des grues à volée horizontale et l'affichage synoptique du résultat (French, Romanian summary)	35
M. ADAM and A. BARABOI, Controlling Dynamic Characteristics of Commutation Electric Equipment (English, Romanian summary)	41
E. HNATIUC, Sur le calcul optimal des électro-aimants à courant alternatif de type U-I munis avec des bagues de déphasage (French, Romanian summary)	45
V. MUNTEANU, M. LUCANU A. PĂSTRĂVANU and D. TĂRNICERU, New Decoding Procedures for Linear Binary Codes (I) (English, Romanian, summary)	49
P. COTAE and N. D. ALEXANDRU, Spectral Width of MKS Type Signaling Waveforms (English, Romanian summary)	57
M.-S. OPRIŞOR and S.-A. GABA, Mathematical Model of the Subsynchronous Condensers Braking of the Asynchronous Three-Phase Motor (English, Romanian summary)	63
J. IGNAT, Une nouvelle méthode d'identification et localisation d'une mise à terre dans un réseau de moyenne tension ayant le point neutre isolé (French, Romanian summary)	71
GH. CÂRŢINĂ, Sur l'évolution de la consommation de l'énergie électrique (French, Romanian summary)	77
BOOK REVIEWS	81

BULETINUL INSTITUTULUI POLITEHNIC IAȘI
Tomul XXXVII (XLI), Fasc. 1—4, 1991
Secția III
ELECTROTEHNICĂ. ENERGETICĂ. ELECTRONICĂ

Acad. Prof. Dr. DIMITRIE I. MANGERON
(1906—1991)

Professor Dr. Dimitrie I. Mangeron, member of the Romanian Academy, a leading figure of the romanian science and high technical teaching, was born in 15/28 November 1906, in Kishinev. After brilliant studies, he graduated in Jassy the Facuity of Sciences of the University „Al. I. Cuza". Then, after two years of more studies in Italy, he became doctor in mathematical sciences, with a remarkable work, devoted to some nonlinear differential equations with partial derivatives. These researches contain in germ what will be later called as „Mangeron equations".

Returned at home, between 1935—1941, he covers all the university ranks, until that of professor. Since 1940 and for all his life, professor Mangeron works with an outstanding passion at the Polytechnic Institute „Gh. Asachi" in Jassy, rapidly becoming a great scientist and a great teacher. He delivered courses on mathematics, but

especially he remains the fascinating professor in theoretical mechanics, for countless series of students, from all the faculties. He enchanted his audience through a wide intellectual opening, striking the listener imagination by a great capacity to penetrate with him in the heart of the more actual scientific problems. For a long time he worked hardly as head of the Departments of Theoretical Mechanics and Mathematics.

During more than sixty years, until the last moment, professor M a n g e r o n was an almost unique example of devotion in the service of teaching and science. With an contaminating enthusiasm and outstanding generosity, he created well-known schools of scientific researches. Open minded and exceptionally gifted, he contributed with remarkable and lasting results to the progress of mathematics and of theoretical and applied mechanics. Through most than 600 papers, written alone or with coworkers from 25 countries, professor Dimitrie M a n g e r o n solved a great number of sharp scientific problems, very impressive due to their up-to-dateness, widtness and variety. He became very soon known and recognized by the international scientific community, who inceasingly asked him to present conferences, to deliver courses or to guide doctoral dissertations, all over the world.

Through his scientific researches, professor M a n g e r o n put his mark on various fields of great interest, such as analytical mechanics, mechanics of vibration, theory of mechanisms and machines, optimal control in engineering, biomechanics, theory and applications of differential equations with partial derivatives, variational methods, history of mathematics and mechanics. Continuing his remarkable achievements, notorious specialists have introduced them in the international scientific circuit, as „Mangeron polyvibrating equations", „Mangeron operators", „Mangeron generalized equations in analytical mechanics", etc.

Scientist of enciclopedic style, profesor Dimitrie M a n g e r o n kept up a vast correspondence with an incredible number of scientists, editorial staffs of reviews, universities and research institutes from all the continents. Simultaneously, he spend incessantly much energy to spead over the world our scientific achievements and to keep up a lively exchange of scientific information. As nobody else, profesor M a n g e r o n contributed to make known the Polytechnic Institute „Gh. Asachi". He was elected member of numerous scientific societies, academies, editorial staffs of prestigious international journals from Europe, Nord and South America and Asia.

One great achievement of professor M a n g e r o n remains the foundation, in 1946, of the now well known „Bulletin of the Polytechnic Institute «Gh. Asachi», Jassy". He succeeded in securing it a high scientific level, who attracted very large contributions of numerous remarkable scientists from all geographical regions. Owing to professor M a n g e r o n, our „Bulletin" has acquired a high international prestige and

a strong capacity to penetrate in the international scientific media, on the basis of material exchange with a wide range of similar issues.

Professor Dimitrie I. Mangeron, member of the Romanian Academy, is definitively entered into the first ranks of our great scientists and teachers of this century. He remains alive in the memory of all who know him and cannot forget his largely open mind to universality, his inexhaustible desire to know and skill to spread knowledge, his unquestionable competence and profound humanism.

Editorial Staff of the
„Bulletin of the Polytechnic
Institute «Gh. Asachi», Jassy"

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Tomul XXXVII (XLI) Fasc. 1—4, 1991
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ELECTROTEHNICĂ. ENERGETICĂ. ELECTRONICĂ

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SUR LA RÉALISABILITÉ DE LA FERRORÉSONANCE DANS LES CIRCUITS NONLINÉAIRES INERTIAUX EN RÉGIME PERMANENT HARMONIQUE

PAR

HUGO ROSMAN

1. Nous avons effectué, dans une Note précédente [1], une première tentative de déterminer les conditions de la réalisabilité du régime de ferrorésonance dans les circuits nonlinéaires inertiaux fonctionnant en régime permanent harmonique. À savoir, on a établi les conditions qui assûrent la réalisation d'un tel régime à la porte d'accès d'un uniporte nonlinéaire inertial alimenté par un uniporte de même nonlinéaire, inertiel, alimenté à son tour par un générateur de tension harmonique. Il faut souligner que l'investigation du problème de la réalisabilité de la ferrorésonance dans des circuits nonlinéaires noninertiaux fait l'objet de nombreuses recherches, parmi lesquelles celles effectuées par S. Mitrea et A. Adăscăliţei [2] ... [4] qui ont abouti à des conclusions très intéressantes.

Le but de cette Note est d'étudier d'une façon plus approfondie que dans celle précédente [1] le problème de la réalisabilité des régimes de ferrorésonance dans les circuits nonlinéaires inertiaux en régime permanent harmonique, utilisant un modèle physique plus simple, capable, toutefois, à fournir des informations supplémentaires, qui presentent de l'intérêt du point de vue pratique.

2. On considère le circuit passif représenté dans la Fig. 1, où tant la bobine que le condensateur sont des éléments nonlinéaires inertiaux. Si les caractéristiques de ces élémentes sont polynômiales ayant la forme [5] :

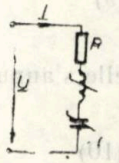

Fig. 1

$$\psi = b_1 i - b_3 i^3, \quad u = c_1 q + c_3 q^3, \tag{1}$$

avec

$$b_1 > 0, \; b_3 > 0 \; ; \; c_1 > 0, \; c_3 > 0, \tag{2}$$

on obtient, pour les réactances de la bobine et du condensateur, les expressions ci-dessous :

$$X_L = b_1 \omega - \frac{3}{4} b_3 \omega I_m^2, \; X_C = \frac{c_1}{\omega} + \frac{3 c_3}{4 \omega^3} I_m^2, \tag{3}$$

où I_m représente l'amplitude du courant, avec la restriction

$$I_m < I_{m0} = \frac{2}{\sqrt{3}} \sqrt{\frac{b_1}{b_3}} \tag{4}$$

afin que X_L soit positif.

En régime permanent harmonique la ferrorésonance est réalisée si :

$$X = X_L - X_C = b_1\omega - \frac{c_1}{\omega} - \frac{3}{4}\left(b_3\omega + \frac{c_3}{\omega^3}\right) I_m^2 = 0, \tag{5}$$

c'est-à-dire si

$$I_m = I_{mr} = \frac{3}{\sqrt{2}} \sqrt{\frac{b_1\omega^4 - c_1\omega^2}{b_3\omega^4 + c_3}}. \tag{6}$$

En vue de la réalisabilité d'un tel régime il est nécessaire que

$$\omega > \omega_0 = \sqrt{\frac{c_1}{b_1}} \tag{7}$$

et

$$I_{mr} < I_{m0}. \tag{8}$$

Si l'on compare les relations (4) et (6) on peut constater que l'inégalité (8) est toujours satisfaite. Il s'ensuit que, dans les conditions qui assûrent la réalisabilité du regime de ferrorésonance y interviennent seulement les coefficients b_1, c_1, des termes linéaires des caractéristiques des éléments nonlinéaires (inertiaux).

Pour un circuit nonlinéaire, inertiel, donné, du type représenté dans la Fig. 1, il résulte, compte tenu des observations ci-dessus, que *pour une certaine valeur de la fréquence du signal excitation, vue la restriction* (7), *il y a une seule valeur de l'amplitude de ce signal* (I_m ou $U_m = RI_m$) *qui assûre la réalisation de la ferrorésonance en régime permanent harmonique.*

3. On peut considérer l'amplitude du courant, I_{mr}, dans le régime de ferrorésonance, comme une fonction de la fréquence (v. (6)). Sa dérivée est :

$$\frac{dI_{mr}}{d\omega} = \frac{2\omega(b_3c_1\omega^4 + 2b_1c_3\omega^2 - c_1c_3)}{\sqrt{3}(b_3\omega^4 + c_3)^{3/2}}; \tag{9}$$

elle s'annulle pour la valeur réelle et positive de la pulsation :

$$\omega_m = \sqrt{\frac{-b_1c_3 + \sqrt{b_1^2c_3^2 + b_3c_1^2c_3}}{b_3c_1}}. \tag{10}$$

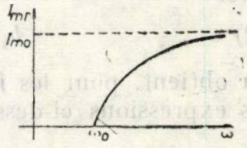

Fig. 2

Si l'on effectue les calculs en peut établir aisément que (v. (7) et (10)) :

$$\omega_m < \omega_0 \tag{11}$$

et par conséquent le graphique $I_{mr}(\omega)$ a la forme représentée dans la Fig. 2.

4. Dans les régimes de ferrorésonance

$$X_{Lr} = X_{Cr} = X_r = \omega\,\frac{b_1c_3 + b_3c_1\omega^2}{c_3 + b_3\omega^4}, \tag{12}$$

de sorte qu'on peut considérer la réactance comme une fonction de la pulsation, ω. Sa dérivée par rapport à ω est :

$$\frac{dX_r}{d\omega} = \frac{-b_3^2 c_1 \omega^6 - 3b_1 b_3 c_3 \omega^4 + 3b_3 c_1 c_3 \omega^2 + b_1 c_3^2}{(b_3\omega^4 + c_3)^2}, \tag{13}$$

qui s'a nulle lorsque

$$-b_3^2 c_1 \Omega^3 - 3b_1 b_3 c_3 \Omega^2 + 3b_3 c_1 c_3 \Omega + b_1 c_3^2 = 0, \quad (\Omega \overset{\Delta}{=} \omega^2). \tag{14}$$

Cette équation a une seule racine réelle et positive ; si l'on tient compte, en plus, que $X_r(0)=0$ et $\lim_{\omega\to\infty} X_r(\omega)=0$, le graphique a la forme représentée dans la Fig. 3. Évidemment, en vue de la réalisabilité du circuit il faut considérer seulement l'arc de courbe qui correspond aux pulsations $\omega_r > \omega_0$. Il est assez difficile à préciser, dans le cas général, quel est le rapport entre les pulsations ω_m et ω_0. On peut observer que

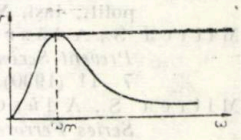

Fig. 3

$$X_r(\omega_0) = \sqrt{b_1 c_1}, \tag{15}$$

valeur qui dépend seulement des coefficients des termes linéaires des caractéristiques de la bobine et du condensateur.

5. Les deux cas particuliers ci-dessous présentent un intérêt spécial :

a) Bobine nonlinéaire et condensateur linéaire ($c_1 = 1/C$, $c_3 = 0$). La relation (6) devient

$$I_{mr} = \frac{2}{\sqrt{3b_3}} \sqrt{b_1 - \frac{1}{\omega^2 C}} \tag{16}$$

et par conséquent

$$\omega_0 = \frac{1}{\sqrt{b_1 C}}; \tag{17}$$

le graphique $I_{mr}(\omega)$ est analogue à celui représenté dans la Fig. 2.

b) Bobine linéaire et condensateur nonlinéaire ($L_1 = L$, $b_3 = 0$). Dan cette situation :

$$I_{mr} = \frac{2\omega}{\sqrt{3c_3}} \sqrt{L\omega^2 - c_1}, \tag{18}$$

de sorte que

$$\omega_0 = \sqrt{\frac{c_1}{L}}. \tag{19}$$

6. Les résultats obtenus ci-dessus correspondent, évidemment, au cas de la ferrorésonance série. On peut aboutir à des conclusions semblables dans le cas de la ferrorésonance parallèle.

7. Les aspects relevés dans cette Note sont de nature à compléter les informations trouvées dans la littérature (par ex. [6]) concernant la ferrorésonance dans les circuits nonlinéaires inertiaux.

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Institut Polytechnique, Jassy,
Chaire d'Electrotechnique

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ASUPRA REALIZABILITĂȚII FERREZONANȚEI ÎN CIRCUITE NELINIARE INERȚIALE ÎN REGIM PERMANENT ARMONIC

(Rezumat)

Se stabilesc unele relații referitoare la amplitudinea curentului de alimentare și frecvența semnalului excitație necesar a fi satisfăcute în vederea realizării regimului de ferorezonanță la poarta unui uniport neliniar inerțial.

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D.C.621.3.014.8

UPON THE DETERMINATION OF CAPACITANCE TO PREVENT SERIES FERRORESONANCE

BY

SAVEL MITREA and ADRIAN ADĂSCĂLIŢEI

1. Introduction

Series ferroresonance is an undesired phenomenon occuring especially in the single-phase switching of unloaded three-phase distribution transformer banks [1], [2]. As shown in [3], the piecewise-linear model of series ferroresonant circuit, based on the magnetic characteristic shown in Fig. 1, is an effective representation of the real circuit. The above characteristic leads to a series ferroresonant circuit as represented in Fig. 2.

Ferroresonance occurs in distribution systems if capacitance, i. e. the cable length, exceeds a certain value. In order to prevent it special steps are recommended in the literature. It is known that input—output amplitude characteristic of the circuit resembles a parabola presenting a minimum input voltage (V_{min}) below which ferroresonance cannot be sustained. V_{min} is a function of ψ_s, I_m, R, L, C, of which, only C is variable. Once the transformer parameters

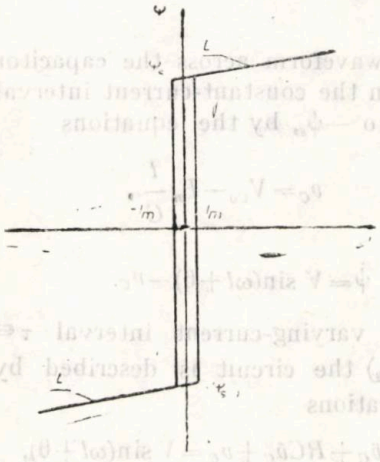

Fig. 1

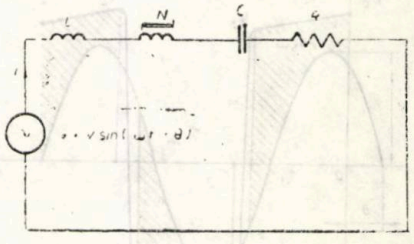

Fig. 2

are known the value of C at which V_{min} equals the nominal voltage is found on the curve $V_{min}=f(C)$. The curve can be obtained by repeatedly calculating input—output amplitude characteristic for varying C. To determine steady-state response one usually resorts to simulation. It is known that in the cases where long-term computer simulation is necessary the results are of questionable validity. Numerical integration to find solution of piecewise-linear models, as the one employed in this paper, involves both truncation and round-off errors [4]. In order to minimize errors one has to use exact analytical formulas which hold in each linear region in conjunction with long-word length computers. However, there are cases where periodic state is not obtainable [5] due to sensitive dependence on initial conditions.

In order to overcome the difficulties mentioned above one has to find those initial conditions that suppress the transient. To this purpose one can resort to the so called shooting method [6] but it involves a lot of guess on initial conditions. Another approach to the problem consists in solving the steady-state equations, readily obtainable once the above mentioned exact analytical formulas are available. This approach does not suffer from the drawbacks encountered in simulation but solving the steady-state equations is not trivial if one sees them as transcendental [3]. The authors of the present paper have found that the steady-state equations can be written in a linear form. Beside the fact that solving the equations is now trivial, the linear form permits the obtaining of an expression for V_{min} on the amplitude characteristic. The linear form of equations and the expression for V_{min} constitute the main contributions of this paper to the study of ferroresonance in distribution systems. If simulation is used the approach presented in this paper constitutes a safe means to check the validity of results if there are reasons to think that they are questionable. We mention that writing the equations in a linear form is not restricted to the circuit considered in this paper. The key point is to single out the first two unknowns in the form shown in the next section.

2. Determination of the Expression for Minimum Voltage to Sustain Series Ferroresonance

Fig. 3 shows the steady-state voltage waveform across the capacitor as presented in [3]. The circuit is described on the constant-current interval $\tau \in (0, \tau_1)$, as flux linkage ψ changes from ψ_s to $-\psi_s$, by the equations

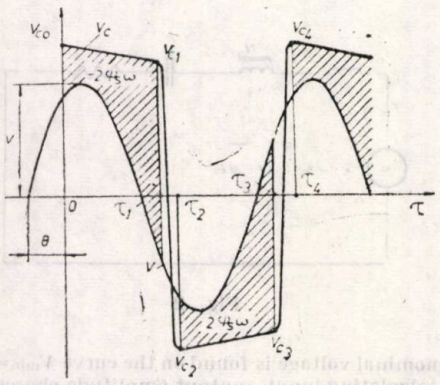

Fig. 3

$$(1) \qquad v_C = V_{C0} - I_m \frac{t}{C}.$$

$$(2) \qquad \dot{\psi} = V \sin(\omega t + \theta) - v_C.$$

On the varying-current interval $\tau \in (\tau_1, \tau_2)$ the circuit is described by the equations

$$(3) \quad LC\ddot{v}_C + RC\dot{v}_C + v_C = V \sin(\omega t + \theta),$$

$$(4) \qquad i = C\dot{v}_C.$$

By using the adimensional variable $\omega t = \tau$ and considering $\psi_0 = \psi_s$ at $\tau = 0$, the expressions $\psi(\tau)$, $v_C(\tau)$, $i(\tau)$ are found by means of linear circuit theory. It should be noted that if errors are to be minimized in simulating the circuit (§ 1) the above expressions must be derived anyhow. By setting the periodicity conditions $V_{C2} = -V_{C0}$ and $\tau_2 = \pi$, one obtains the steady-state equat-

ions that describe the response wareform. By appropriate manipulations the equations can be written as

$$\sum_{j=1}^{3} a_{ij}x_i = b_i, \qquad (i=1, 2, 3), \tag{5}$$

where : $x_1 = V \sin\theta$, $x_2 = V\cos\theta$, $x_3 = V_{C0}$, $a_{ij} = f_{ij}(\tau_1, R, L, C, \omega)$, $b_i = f_i(I_m, \psi_s, R, L, C, \omega)$. The expressions for a_{ij} and b_i are given in Appendix.

By assigning values to the parameter τ_1 on the interval $\tau_1 \in (0, \pi)$, exact values for x_i can be found from (5). The solutions are valid if the condition $v_C(\tau) > v(\tau)$ is fulfilled for all $\tau \in (0, \tau_1)$. There is no need to carry out the validity test if one is interested in V_{min} only ; the above condition is always fulfilled in this case. Given C, exact value for V_{min} is found on the input–output amplitude characteristic $V = f(V_{C0})$ obtained by repeatedly solving (5), τ_1 taking values in the interval mentioned above. By applying this procedure for different values of C, the curve $V_{min} = f(C)$ can be plotted in order to find C_{max} corresponding to nominal voltage $V_{nom} = \omega\psi_s$.

Although solving linear algebraic equations in order to plot amplitude characteristics is much easier than in the case they were transcendental, a formula to give in a strightforward manner the relationship $V_{min} = f(C)$ is desirable. An exact formula for V_{min} cannot be derived because the value of τ_1 corresponding to V_{min} cannot be obtained explicitly (s. expressions for a_{ij} in Appendix). Yet, an almost exact formula is obtained from (5) if the ength of constant-current interval takes the value

$$\tau_1 = \pi\left(1 - \frac{\omega}{\Omega}\right) \overset{\Delta}{=} \alpha, \tag{6}$$

where $\Omega = \sqrt{1/LC - [R/(2L)^2]}$ is the damped natural frequency of the loop, consisting of the elements R, L, C of the circuit in Fig. 1. Taking into account that (5) is a linear system of algebraic equations, we can write the expression

$$V_{min} = \frac{1}{|\Delta|}\sqrt{\Delta_1^2 + \Delta_2^2}\,. \tag{7}$$

The approximate expressions for a_{ij} are given in Appendix, for convenience.

Although (7) gives $V_{min} = f(C)$, the capacitance cannot be expressed in terms of V_{min} so that we still have to construct the above function in order to find C for which $V_{min} \geqslant V_{nom}$. Besides, as it results from [3], determination of parameters R and L encounters some difficulties. In what follows we set $R = 0$ and L very small in the approximate expressions for a_{ij} in Appendix. We get

$$\theta = \frac{\pi}{2}, \tag{8}$$

$$V_{min} = \frac{\pi I_m}{2\omega C}. \tag{9}$$

$$V_{C0} = \frac{1}{\pi}\left(2V_{nom} + \frac{\pi^2 I_m}{2\omega C}\right). \tag{10}$$

Expression (9) shows that V_{min} increases as C decreases. This is true of the exact value for V_{min} as well. It also results from (9) that in order to rule

out ferroresonance, i. e. $V_{min} > V_{nom}$, the following condition has to be fulfilled

$$C < \frac{\pi I_m}{2\omega V_{nom}}. \tag{11}$$

3. Application

By taking into account the best fit parameters for a distribution transformer [3] and using the exact expressions for a_{ij} in (5), we have found $C_{exact} = 7.1$ μF, corresponding to $V_{min} = V_{nom} = \omega\psi_s$. By using the above mentioned best fit parameters and applying approximate expression (7) we have found $C_{approx.(1)} = 6.96$ μF, which differs by 1.97% from the exact value for C. By using approximate expression (11), that ignores parameters R and L, we have found $C_{approx.(2)} = 6.09$ μF, which differs by 14.2% from the exact value for C. If compared to the exact value this last result brings out the role the magnetizing current I_m plays in applications of this nature. It must be stressed that, in contrast to other methods applied for the study of ferroresonance, the piecewise-linear approach permits the taking into account the magnetizing current without increasing complexity of the analysis, provided that the validity condition mentioned in § 2 is fulfilled.

4. Conclusions

Due to the linear form in which steady-state equations are written, determination of the value of capacitance beyond which series ferroresonance can develop is substantially simplified. Moreover, the linear form permits the obtaining of expression (7) to give almost exact value for the minimum applied voltage to sustain series ferroresonance. To our best knowledge, this is the first time an expression for V_{min} has been derived. If this expression is considered to be involved and/or the dynamic parameters R, L difficult to determine approximate expression (11) is obtained by ignoring the above parameters. If accuracy of the results so obtained are considered as unacceptable, capacitance obtained by using (11) can be considered as a start in applying expression (7) or the exact procedure. However, if the value of C given by (11) is considered, the user is on the safe side since this value is smaller than the almost exact and the exact one.

Appendix

R, L, C — circuit parameters ;

$R_u = 2\sqrt{L/C}$ — critical resistance of RLC loop ;

ω — angular frequency of sine voltage source ;

$\Omega = \sqrt{1/(LC) - [R/(2L)]^2}$ — damped natural frequency of RLC loop ;

$n = \Omega/\omega$;

$\delta = \pi R/(\Omega L)$ — logarithmic decrement of amplitude of underdamped natural response of RLC loop ;

$R_n = R/R_{cr} = \delta/\sqrt{4\pi^2+\delta^2}$ – normalized resistance ;

$X_n = X_C/R_{cr} = n\sqrt{4\pi^2+\delta^2}/(4\pi)$ – normalized capacitive reactance ;

$Z_n = \sqrt{R_n^2+[1/(4X_n)-X_n]^2}$ – modulus of normalized impedance ;

$\varphi = \tan^{-1}\{[1/(4X_n)-X_n]/R_n\}$ – argument of complex impedance ;

$V = \tan^{-1}(2\pi/\delta)$;

$\gamma_{nom} = \omega\psi_s$ – nominal voltage ;

I_m – magnetizing current ;

$a_{11}, \ldots, a_{33}$ – coefficients in eqs. (5) ;

$a_{11} = \sin\tau_1$; $a_{12} = 1-\cos\tau_1$; $a_{13} = -\tau_1$;

$$a_{21} = X_n\{\sin\gamma\sin\varphi\exp[n\delta(\pi-\tau_1)]/(2\pi)] - \sin[(n(\pi-\tau_1)+\gamma]\sin(\tau_1-\varphi)\} - \frac{1}{2}\sin[n(\pi-\tau_1)]\cos(\tau_1-\varphi) ;$$

$$a_{22} = X_n\{\sin\gamma\cos\varphi\exp[n\delta(\pi-\tau_1)/(2\pi)] + \sin[n(\pi-\tau_1)+\gamma]\cos(\tau_1-\varphi))\} - \frac{1}{2}\sin[n(\pi-\tau_1)]\sin(\tau_1-\varphi) ;$$

$$a_{23} = Z_n\{\sin[n(\pi-\tau_1)+\gamma] + \sin\gamma\exp[n\delta(\pi-\tau_1]/(2\pi)]\},$$

$$a_{31} = -\sin\gamma\cos\varphi\exp[n\delta(\pi-\tau_1)/(2\pi)] + \sin[n(\pi-\tau_1)-\gamma]\cos(\tau_1-\varphi) + 2X_n\sin[n(\pi-\tau_1)]\sin(\tau_1-\varphi) ;$$

$$a_{32} = \sin\gamma\sin\varphi\exp[n\delta(\pi-\tau_1)/(2\pi)] + \sin[n(\pi-\tau_1)-\gamma]\sin(\tau_1-\varphi) - 2X_n\sin[n(\pi-\tau_1)]\cos(\tau_1-\varphi) ;$$

$$a_{33} = -2Z_n\sin[n(\pi-\tau_1)] ;$$

$b_1, \ldots, b_3$ – right-hand terms of (5) ;

$$b_1 = -2V_{nom} - X_n R_C I_m \tau_1^2/2 ;$$

$$b_2 = Z_n R_C I_m\left\{X_n\tau_1\sin[n(\pi-\tau_1)+\gamma] + \frac{1}{2}\sin[n(\pi-\tau_1)]\right\} ;$$

$$b_3 = Z_n R_C I_m\{\sin\gamma\exp[n\delta(\pi-\tau_1)/(2\pi)] - \sin[n(\pi-\tau_1)-\gamma]\} - 2X_n\tau_1\sin[n(\pi-\tau_1)] ;$$

$a'_{11}, \ldots, a'_{33}$ – coefficients of (5) after substitution ;

$\tau'_{11} \overset{\Delta}{=} \alpha = \pi(1-1/n)$;

$a'_{11} = \sin\alpha$; $a'_{12} = 1-\cos\alpha$; $a'_{13} = -\alpha$; $a'_{21} = \sin\varphi\exp(\delta/2) + \sin(\alpha-\varphi)$;

$a'_{22} = \cos\varphi\exp(\delta/2) - \cos(\alpha-\varphi)$; $a'_{23} = [\exp(\delta/2)-1]Z_n/X_n$;

$a'_{31} = -\cos\varphi\exp(\delta/2) + \cos(\alpha-\varphi)$; $a'_{32} = \sin\varphi\exp(\delta/2) + \sin(\alpha-\varphi)$; $a'_{33} = 0$;

$b'_1, \ldots, b'_3$ – right-hand terms of (5) after substitution $\tau'_1 \overset{\Delta}{=} \alpha = \pi(1-1/n)$;

$b'_1 = -2V_{nom} - X_n R_C I_m \alpha^2/2$; $b'_2 = -Z_n R_r I_m \alpha$; $b'_3 = Z_n R_C I_m[\exp(\delta/2)-1]$.

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ASUPRA DETERMINĂRII CAPACITĂȚII PENTRU EVITAREA FEROREZONANȚEI SERIE

(Rezumat)

În scopul analizei circuitului ferorezonant se utilizează liniarizarea pe porțiuni a caracteristicii de magnetizare. Apariția ferorezonanței este posibilă dacă amplitudinea tensiunii aplicate depășește o anumită valoare, V_{min}. Datorită caracterului transcendent al ecuațiilor care descriu regimul periodic nu se poate obține o expresie analitică pentru a calcula valoarea exactă a tensiunii V_{min}. Se arată că aceste ecuații pot fi scrise sub o formă liniară ceea ce permite obținerea unei expresii pentru V_{min} care conduce la valori aproape exacte. Forma liniară a ecuațiilor permite determinarea valorii exacte, aproape exacte și aproximative a valorii capacității dincolo de care ferorezonanța este posibilă atunci cînd se aplică tensiunea nominală a transformatorului de distribuție.

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A REPRESENTATION OF FINITE DISTRIBUTED SYSTEMS

BY
IRINEL CASIAN

1. Introduction

An electrical system is called as distributed if it is represented by a system of linear partial differential equations, of first order in the spatial dimension x and of arbitrary order in the time dimension t. The equations for a such system are :

$$\begin{aligned}
\frac{\partial q_1}{\partial x} &= \left(f_{11}^0 + f_{11}^1 \frac{\partial}{\partial t} + \ldots + f_{11}^r \frac{\partial^r}{\partial t^r}\right) q_1 + \ldots + \\
&+ \left(f_{1n}^0 + f_{1n}^1 \frac{\partial}{\partial t} + \ldots + f_{1n}^r \frac{\partial^r}{\partial t^r}\right) q_n + u_1(t, x), \\
&\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \\
\frac{\partial q_n}{\partial x} &= \left(f_{n1}^0 + f_{n1}^1 \frac{\partial}{\partial t} + \ldots + f_{n1}^r \frac{\partial^r}{\partial t^r}\right) q_1 + \ldots + \\
&+ \left(f_{nn}^0 + f_{nn}^1 \frac{\partial}{\partial t} + \ldots + f_{nn}^r \frac{\partial^r}{\partial t^r}\right) q_n + u_n(t, x),
\end{aligned} \tag{1}$$

where $q_i = q_i(t, x)$ are voltages or currents and $u_i(t, x)$ — forcing terms.

The transmission line and all its analogs fall into this category, as do many other systems. In matrix form (1) reads :

$$\frac{\partial [q]}{\partial x} = [F_0][q] + [F_1] \frac{\partial [q]}{\partial t} + [F_2] \frac{\partial^2 [q]}{\partial t^2} + \ldots + [F_r] \frac{\partial^r [q]}{\partial t^r} + [u]. \tag{2}$$

The matrices $[F_i] = [f_{jk}]$ may be constants or, more generally, functions of x. Hereafter we shall terminate the series in (2) with the third matrix $[F_2]$, since this is sufficient for most systems.

If it is assumed that the solution of (1) exists and possesses a Laplace transform with respect to t, (2) can be transformed into :

$$\frac{d[Q]}{dx} = ([F_0] + [F_1]s + [F_2]s^2)[Q] + [\bar{U}](s, x), \tag{3}$$

where

$$[\bar{U}](s, x)=[U](s, x)-[F_1][q](0, x)-[F_2]([\dot{q}](0, x)+s[q](0, x)$$

and $[U](s, x)=\mathcal{L}([u](t, x))$, ($\mathcal{L}$ — Laplace transform).

In acordance with the theory of differential ordinar equations, the solution of (3) can be expressed by :

$$[Q](s, x)=[H](s, x-x_0)[Q](s, x_0)+\int_{x_0}^{x}[H](s, x-\xi)[\bar{U}](s, \xi)\mathrm{d}\xi, \tag{4}$$

where $[H](s, x)$ is the fundamental matrix for the homogeneous equation from (3). In particular, when $[F_0]$, $[F_1]$ and $[F_2]$ are constant matrices, the fundamental matrix is :

$$[H](s, x)=\exp\{([F_0]+[F_1]s+[F_2]s^2)x\}=[I]+([F_0]+[F_1]s+$$
$$+[F_2]s^2)x+([F_0]+[F_1]s+[F_2]s^2)\frac{x^2}{2!}+\ldots \tag{5}$$

From expansion of the series defining $[H](s, x)$, it will be found that each element of $[H](s, x)$ is a power series in x whose coefficients are polynomials in s. Since the series defining an exponential converges for all finite values of x is any finite (complex) number, it is thus necessary for each element in the matrix $[H](s, x)$ to be analytical for all finite value of s for all finite values at x [1].

2. The Equations for Finite Systems

The basic techniques illustrated by the electrical transmission line [2], and its analogs, can be applied to the more general system described by (3).

The dynamic behaviour of a physical distributed system depends not only on the partial differential equations governing its internal behaviour, but also on what happens at the ends, or terminals.

Consider the general system :

$$\frac{\partial[q]}{\partial x}=[F_0][q]+[F_1]\frac{\partial[q]}{\partial t}+[F_2]\frac{\partial[q]}{\partial t^2}+[u](t, x), \tag{6}$$

with the general frequency-domain solution (4).

If the system is of order n in space (that i $[q]$ and $[Q]$ are n-vectors), the equation (4) gives n relations between the n quantities at a point x and n quantities at a point x_0.

Now suppose that the sys em is of finit length, $0\leqslant x\leqslant L$; then we have, at $x=L$:

$$[Q](s, L)=[H](s, L)[Q](s, 0)+\int_0^L [H](s\ L-\xi)[\bar{U}](s, \xi)\mathrm{d}\xi. \tag{7}$$

To determine $[Q](s, L)$ and $[Q](s, 0)$ (and hence $[Q](s, x)$ for any $x \in [0, L]$), n terminal relations must be specified. These relations can generally be put in the form of n independent ordinary differential equations:

$$\begin{aligned} &m_{11}(D)q_1(t, 0) + \ldots + m_{1n}(D)q_n(t, 0) + n_{11}(D)q_1(t, L) + \ldots + \\ &\quad + n_{1n}(D)q_n(t, L) = z_1(t), \\ &\qquad \cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots \\ &m_{n1}(D)q_1(t, 0) + \ldots + m_{nn}(D)q_n(t, 0) + n_{n1}(D)q_1(t, L) + \ldots + \\ &\quad + n_{nn}(D)q_n(t, L) = z_n(t), \end{aligned} \tag{8}$$

where $m_{ij}(D)$ and $n_{ij}(D)$ are linear differential operators, and $z_i(t)$ are external forcing functions (the physical inputs or linear combinations thereof).

In vector-matrix form, (8) can be written:

$$[M](D)[q](t, 0) + [N](D)[q](t, L) = [z](t). \tag{9}$$

Laplace transformation of (9) results in the frequency-domain expression for the terminal relations:

$$[M](s)[Q](s, 0) + [N](s)[Q](s, L) = [\bar{Z}](s), \tag{10}$$

where $[\bar{Z}](s)$ is the sum of the Laplace transform of $[z](t)$ and any initial condition terms arising in the transformation of (9), transposed to the right-hand side. The termination is said to be *homogeneous* if $[\bar{Z}](s) = 0$.

Note that each of the differential equations of (8) can involve quantities at both ends of the distributed system. This is the type of termination characteristic of a feedback system. When no feedback is involved, (10) would become two separate equations:

$$[A](s)[Q](s, 0) = [\bar{Z}_1](s), \quad [B](s)[Q](s, L) = [\bar{Z}_2](s),$$

where the first represents n_1 independent equations and the second, $n - n_1$ independent ones.

In order to obtain a complete representation of the system governed by (4) and (10), we substitute the former into the latter with $x_0 = 0$ and $x = L$; it result:

$$[M](s) + [N](s)[H](s, L)[Q](s, 0) = [V](s),$$

where

$$[V](s) = [\bar{Z}](s) - [N](s) \int_0^L [H](s, L - \xi)[\bar{U}](s, \xi)\, d\xi. \tag{11}$$

If the relations (8) are independent (which is a requirement of a fully specified system), $[M](s) + [N](s)[H](s, L)$ is of rank n and has an inverse. Hence we can write:

$$[Q](s, 0) = \{[M](s) + [N](s)[H](s, L)\}^{-1}[V](s). \tag{12}$$

More generally, using (4) with $x_0=0$ we obtain :

$$[Q](s, x)=[H](s, x)\{[M](s)+[N](s)[H](s, L)\}^{-1}[V](s)+$$

$$+\int_0^x [H](s, x-\xi)[\bar{U}](s, \xi)\mathrm{d}\xi. \tag{13}$$

3. Concluding Remarks

The formal expression (13) is convenient for theoretical reasons and as a method of avoiding computational errors. The following result shows the utility of this general formalism.

If there are no distributed forcing terms and no initial conditions, (13) can be written :

$$[Q](s, x)=[H](s, x)\{[M](s)+[N](s)[H](s, L)\}^{-1}[Z](s), \tag{14}$$

where $[Z](s)$ is the Laplace transform of the external forcing terms in (8). Hence the matrix of transfer functions from $[Z](s)$ to $[Q](s, x)$ is given by:

$$[K](s, x)=[H](s, x)\{[M](s)+[N](s)[H](s, L)\}^{-1}.$$

Each term in the matrix $[M](s)+[N](s)[H](s, x)$ is a sum of products of entire functions and is thus an entire function (an entire function of a complex variable s is one which is analytic for all finite value of s). All elements of its adjoint are also entire functions. Likewise, the determinant $|\{[M](s)+[N](s)\}\times \times[H](s, L)|$ is a sum of products of entire functions and is thus an entire function. Consequently, every term of the matrix $[K](s, x)$ is a ratio of entire functions of s. Since the ratio of entire functions is a meromorphic function, [1], it result that the transfer function from any input to any dynamic variable in the distributed system (4), whose terminal relations are given by (8), is a meromorfic function of s. This means that the only singularities which any transfer function may have in the finite s plane are poles.

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O TRATARE A SISTEMELOR CU CONSTANTE DISTRIBUITE

(Rezumat)

În contrast cu teoria care există pentru sistemele guvernate de ecuații diferențiale ordinare, teoria dinamicii sistemelor distribuite este încă incomplet dezvoltată. Din acest motiv s-a încercat o tratare matriceală a sistemelor distribuite, evidențiindu-se în mod firesc noțiunea de reacție și deducându-se o proprietate a funcției de transfer în astfel de sisteme.

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ON PROXIMITY RELAYS WITHOUT CONTACTS

BY

I. SUCHAR and RADU URSULEAN

1. Introduction

This paper presents an application of series ferroresonant circuit with nonlinear coil orthogonally driven by a magnetic field B_c, produced by a permanent magnet which has a linear motion towards the nonlinear coil, as shown in Fig. 1.

Theoretical and experimental studies on ferroresonance in RLC series circuits with nonlinear coil orthogonally driven like the one presented in Fig. 1 were described in [3] ... [5].

2. Theoretical Background

As shown in [1] ... [3], when the induction B_c of the magnetizing field which orthogonally drives the ferromagnetic core is increasing its value, two facts are most important : the gradually linearization of the magnetic cycle and the saturation of the ferromagnetic core, when the magnetic induction B_c of the resultant field has gradually descending values (Fig. 2).

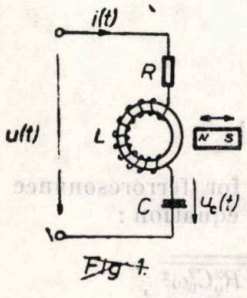

Fig. 1

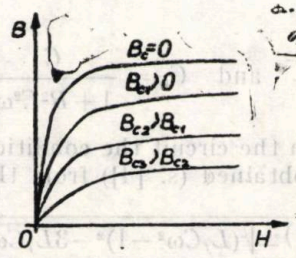

Fig. 2

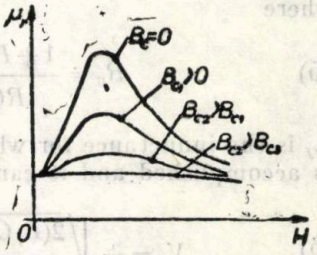

Fig. 3

The magnetization characteristic of the core of the nonlinear coil orthogonally driven like in Fig. 1 can be approximated using a third grade polynomial function :

$$i = b_1\psi + b_3\psi^3, \tag{1}$$

where ψ is the magnetic flux of the coil ; $b_1 > 0$ and $b_3 > 0$ are coefficients whose numerical values are dependent of the magnetic properties of the core.

As B_c is increasing its value, a decrease of the relative magnetic permeability of the ferromagnetic core takes place, and the dependence $\mu_r(H)_{B_c=\text{const.}}$ becames more linear (Fig. 3). Starting from a certain value of the magnetic induction, B_c, the permeability becames independent of the intensity of the main field, H. It means that the reactance of the nonlinear coil (when $B_c=0$) is greater than in the case of an orthogonal premagnetization ($B_c>0$).

It is known that one specific feature of the series ferroresonant circuit with nonlinear coil is that of a relay characteristic, because the current $i(t)$ can have two steady states : one with a high amplitude and another with a low one. Switching from one state to another can be obtained varying the voltage of the power supply $u(t)$, the frequency, or one of the three parameters of the circuit: R, L or C.

What happens when the switching condition for the ferroresonant jump of the amplitude of the current is accomplished can be described using a second order nonlinear differential equation of D u f f i n g type :

$$\frac{d^2v}{d\tau^2} + \frac{1}{RC\omega}\cdot\frac{dv}{d\tau} + \frac{1}{LC\omega^2}v + \frac{b_3\psi_n^2}{LC\omega^2}v^3 = \sin\tau, \tag{2}$$

where

$$\tau = \omega t + \operatorname{arctg} RC\omega\,; \quad \psi_n = \frac{\sqrt{2}U\sqrt{1+R^2C^2\omega^2}}{RC\omega^2}\,; \quad v = \frac{\psi}{\psi_n}. \tag{3}$$

When the circuit in Fig. 1 is powered by an alternating voltage with amplitude $\sqrt{2}U$ and frequency f, both of constant value, the conditions for ferroresonance to occur are :

$$\omega > \frac{1}{\sqrt{L_fC_0}} \text{ and } R_0 > \left|\frac{\sqrt{3}}{C_0\omega - 1/L_f\omega}\right|, \tag{4}$$

where

$$R_0 = \frac{1+R^2C^2\omega^2}{RC^2\omega} \text{ and } C_0 = \frac{C}{1+R^2C^2\omega^2}, \tag{5}$$

L_f is the inductance for which in the circuit the condition for ferroresonance is accomplished and it can be obtained (s. [4]) from the equation :

$$V_f = \pm\sqrt{\frac{2(L_fC\omega^2-1)\pm\sqrt{(L_fC\omega^2-1)^2-3L_fC\omega^2/R_0^2C_0^2\omega^2}}{2{,}25b_3\psi_n^2}}, \tag{6}$$

where V_f is the amplitude of the magnetic flux in the core of the coil; V_f is proportional to the amplitude of the voltage across the coil L_f, when ferroresonance occurs.

The variation of the voltage across the linear capacitor U_c, as a function of the inductivity of the orthogonally premagnetized coil, has also a relay characteristic which is presented in Fig. 4, where the notation:

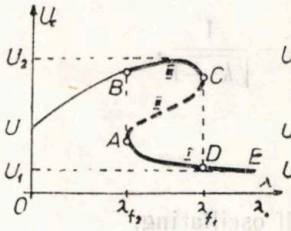

Fig. 4

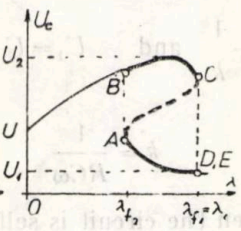

Fig. 5

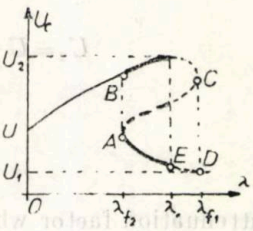

Fig. 6

(7) $$\lambda = LC\omega^2.$$

was utilized.

Thus, when the magnet doesn't magnetize the core of the nonlinear coil, the inductivity is the greatest:

(8) $$\lambda = \lambda_0 = L_0 C\omega^2,$$

where L_0 represents the maximum linear inductivity of the coil when the core isn't magnetized. In this case the amplitude of the voltage U_c is small and is determined by the minimum current that flows in the circuit in the absence of the ferroresonance. When the permanent magnet produces a small premagnetization of the core of the coil, such as $\lambda \in [\lambda_{f_2}, \lambda_0]$, the points of stable functioning are situated on the portion DA of the $U_c(\lambda)$ characteristic.

When the magnet is so close to the coil as the condition $\lambda = \lambda_{f_2}$ is fulfiled, there is a jump to a higher value of the amplitude of the voltage U_c and the functioning point will be situated on the stable branch CB of the characteristic in Fig. 4. The voltage $U_c(\lambda)$ will remain at a high level as long as the parameter λ is somewhere between λ_{f_2} and λ_{f_1}.

If the magnetization is reduced by removing the magnet so that $\lambda = \lambda_{f_1}$, a new jump of the amplitude of U_c will take place from a high value (suitable for the branch CB) to a smaller one (suitable for the branch DA) of the characteristic in Fig. 4.

It must be stressed that the series ferroresonant circuit with nonlinear coil orthogonally driven can be used for a proximity relay without contacts only if in the variation interval of the jumps of the amplitude of U_c from the small value to the big one and viceversa take place, or, to put it another way, only if the $U_c(\lambda)$ function has vertical tangents in two distinct points (A and C, Fig. 4, where the stable branches DA and CB are continued by the unstable branch AC).

The existence of the jumps of U_c in both directions is assured only if $\lambda_0 \geq \lambda_{f1}$ (Figs. 4 and 5) and the conditions (4) are simultaneously fulfiled.

There is one exception from this rule, when $R=\infty$; in this case the characteristic $U_c(\lambda)$ has a single vertical tangent and therefore the jump for U_c is possible only in one way [3]. Of corse in the case when $\lambda_{f_1}>\lambda_0$ theoretically the jump in only one sense can be obtained (Fig. 6).

From Figs. 4 and 5 it can be seen that only the stable branches of the characteristic $U_c(\lambda)$ are of practical uses ; they were drawn with a continuous line. The voltages U_1 and U_2 are given by the following formulas :

$$U_2=U+\frac{1}{k} \quad \text{and} \quad U_1=U-\frac{1}{\sqrt{k^2+1}}, \tag{9}$$

where

$$k=\frac{1}{RC\omega} \tag{10}$$

is the attenuation factor when the circuit is self oscillating.

3. The Elements and the Functioning of the Relay

The basic principle of working of a proximity relay without contacts, as shown in Fig. 7, is the jump in both senses of the voltage drop U_c across the capacitor C, when the core of the nonlinear coil L — the sensitive element of the relay — is orthogonally driven by the magnetic field of a permanent magnet. The magnetic permeability of the core must be high and its magnetic cycle must be rectangular.

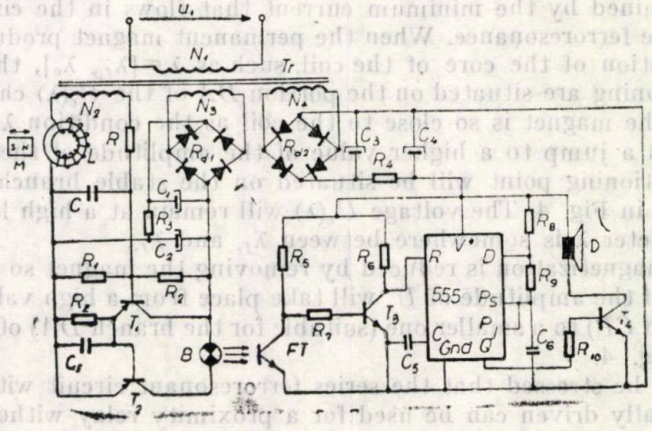

Fig. 7

The schematic diagram of the relay presented in Fig. 7 has two transistors : an emitter follower, T_1, and an amplifier, T_2. The load of T_2 is a lamp, B, for optical signalling. The light of the lamp can drive a phototransistor which permits self oscillating of an astable circuit.

The emitter follower is used for impedance matching, because the *RLC* series ferroresonant circuit has a high otuput impedance and the amplifier has a low input one.

The ferroresonant circuit is powered from a mains supply of industrial frequency by means of the transformer *Tr*. The voltage in the secondary coil of the transformer has a lower amplitude that the one necessary for ferroresonance to occur when the nonlinear coil isn't driven by the orthogonally applied field. The other circuits have their own windings in the secondary of the transformer *Tr* and they are powered by means of the rectifiers Rd_1 and Rd_2.

If the magnet *M* is brought near the nonlinear coil, an orthogonal magnetization of the coil takes place and that is why the inductance *L* lowers its value until ferroresonance occurs. The jump of the amplitude of the current from the small value to a bigger one is the reason of the variation of the voltage drop U_c across the capacitor *C*.

The voltage $u_c(t)$ is applied at the input of the emitter follower T_1 and the signal in the emitter of T_1 switches on the transistor T_2 and therefore the lamp *B*. The base of the phototransistor *FT* is driven by the light of the lamp ans its collector voltage is low ; the transistor T_3 is off and this state permits free oscillation of the astable circuit LM 555. The transistor T_4 with the loudspeaker *D* as load, amplifies the signal from the astable circuit, bringing it to a suitable level for the loudspeaker.

When the magnet *M* is removed there is a reverse jump of the current in the circuit (from a high value of the amplitude to a smaller one). The transistors and the lamp are in this case off ; with no light the phototransistor *FT* is off, T_3 is on, the astable circuit isn't driven, so the sound in the loudspeaker stops.

4. Conclusions

The presented proximity relay signals optically and acoustically when between the permanent magnet *M* and the nonlinear coil is a distance of less than 50 mm. This distance can be adjusted upon request. It depends of the parameters *R*, *L*, *C*, of the level of the voltage in the secondary windings of the transformer, and of the properties of the permanent magnet.

This relay has the following advantages : a) it has no contacts and therefore can be used in harsh environments ; b) the power supply is the industrial main one ; c) the sensitive element has a simple and robust construction and its lifetime is practically unlimited ; d) the electronic circuits don't require neither periodical adjutments nor compensation for the influence of the enviromental factors ; e) precision and safety are independent of temperature and humidity ; f) the sensitive element can be installed at a very high distance from the rest of the circuits, so that the relay can be used for optical and acoustical signalling at high distances.

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ASUPRA UNUI RELEU DE PROXIMITATE FĂRĂ CONTACTE

(Rezumat)

Se propune un nou releu electronic de proximitate, fără contacte, care constituie o aplicaţie a circuitului ferorezonant serie cu bobină neliniară comandată ortogonal, cîmpul de comandă fiind generat de un magnet care se deplasează rectiliniu în raport cu bobina neliniară fixă.

Releul studiat, a cărui soluţie constructivă şi principiu de funcţionare sînt prezentate în brevetul de invenţie [5], se alimentează de la reţeaua industrială, poate fi instalat şi în medii inflamabile cît şi în medii cu pericol de explozie şi este utilizabil în automatizări, la telesemnalizarea optică şi acustică a nivelelor limită admisibile ale lichidelor din rezervoare şi ale materialelor granulare sau pulverulente din silozuri, ca limitator de cursă fără contacte, în robotică etc.

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Secţia III

D.C. 621.317.7.082.7

APPARATUS FOR MEASURING TRIBOELECTRIC ELECTRIFICATION OF INSULATING POWDERS

BY

M. ANTONIU, EMIL VREMERĂ, MARIANA BUZDUGA and L. CHIABURU

1. Introduction

Knowledge of insulating powders electrification capability (flour, coal dust, synthetic resin) has a practical importance due to technologic difficulties (jet deviations, deposition on the tubulation walls) caused by these charges and especially due to fire and explosion that can bring about [1], [2]. The electrostatic charged powders as, for instance, electrostatic coating and xerography, also has a great importance in practical use.

Preoccupied to answer to these needs, the authors elaborated a device that allows electrostatic charge measuring both for pure powders and for mixtures of two powders, with possibility of separating one component in electric field. Measuring range : $10^{-8} \ldots 10^{-11}$ C/g.

2. Electrostatic Charge Measuring of Unmixed Powders

2.1. Measuring Principle

The method used by the authors is based on F a r a d a y's experiment [5]. The device proposed is shown in Fig. 1, where CF_1 is a F a r a d a y cup (a deep cylindrical box of cca. 50 cm³), CE — a screen cage, VEG — electrostatic generating voltmeter. VEG was prefered to an usual electrostatic voltmeter because it allows the potential U_x measurement without galvanic contact with the object.

A small amount of charged powder put in CF_1 causes its potential increase compared to ground level (CE) to U_x. The total powder charge is

$$Q_x = C_p U_x, \tag{1}$$

where C_p is total capacitance of CF_1.

If Q_x is known, specific charge q_x can be calculated, namely :

$$q_x = \frac{Q_x}{m}, \tag{2}$$

where m is the powder mass contained in CF_1.

C_p measurement is also possible by using the capacitor discharge method as follows :

The device is completed as shown in Fig. 2, with : R_0 — a calibrated resistor with a very high value (10^{11} ... 10^{12} Ω) ; U_a — a dc controlled source of 300 ... 500 V and K_a — a very high insulation resistance swich.

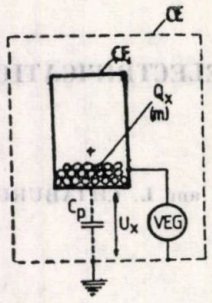

Fig. 1

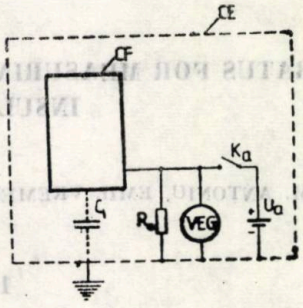

Fig. 2

The cup CF_1 being cleaned, K_a is swiched ON and a value U_a is chosen so that VEG dial reaches the end of the chosen range, E_{01} (end of scale). K_a is then swiched OFF and a chronometer is turned ON simultaneously while VEG indication reaches the value of E_{02} (chosen by the operator).

C_p may be calculated with the relation:

$$C_p = \frac{t}{R_0 \ln \frac{E_{01}}{E_{02}}}. \tag{3}$$

2.2. Electrostatic Generating Voltmeter VEG

The VEG block diagram elaborated by authors is shown in Fig. 3. The rotating electrode *1* (grounded by brush *2* and turned by motor M), together with guard electrode *4* and resistor *R*, form a modulating transducer of electrostatic field, as follows. The unknown voltage U_x is applied to a disc

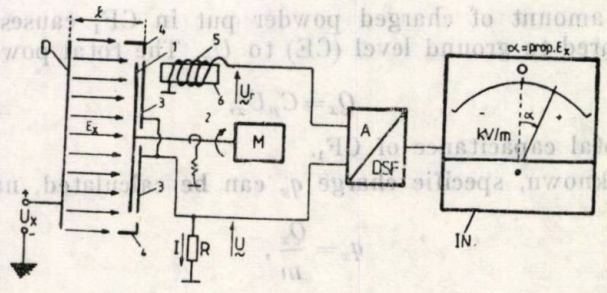

Fig. 3

D ; an uniform electric field exists between the disc and electrode *3* (parallel to D), with the intensity

$$E_x = \frac{U_x}{l}, \tag{4}$$

where l is the distance between electrodes *3* and *1*.

E_x field modulation by electrode *1* generate an alternating current I locked by resistor R, determining a voltage drop proportional to E_x. This voltage, $U\sim$, is first amplified (A) then rectified in a phase sensible detector (DSF) and after that applied to a middle zero magneto-electric instrument (IN), indicating both magnitude (KV/m) and polarity (+ or −) of E_x. The disc D is mechanically and electrically coupled to CF_1.

The reference voltage U_r for DSF is also generated by the same rotating electrode *1* (ferromagnetic) that, together with coil *5* and magnet *6*, form an alternator.

In the variant realized by authors VEG allows E_x measurements from 0.2 to 1 000 KV/m, with a precision higher than 5% (dial end). This domain is shared in seven ranges 1 ; 2.5 ; 10 ; ... ; 1 000 KV/m (for technological details, s. [6]).

2.3. Q_x Measuring Operations

1. CF_1 is short circuited to ground for eliminating any preexisting charge.

2. With CF_1 disconnected from ground, VEG indicates any spurious electric field comming from athmosphere or from electret effect of insulators close to VEG field transducer.

3. Without touching CF_1, the charged powder is put in it and E_x is read on the instrument IN_1 of VEG. The unknown charged is computed with the relation:

$$Q_x = C_p(U_x - U_p) = K_1(E_x - E_p), \tag{5}$$

where U_x and U_p fulfil relation (4) and

$$K_1 = lC_p$$

is the measuring device constant (CF_1—VEG).

4. CF_1 is dismounted, the powder is weighed (m) and q_x is determined from relation (2).

3. Measurements of Toner Powders

From powder mixtures used in electrostatic techniques there will be considered the case of those used in xerographic devices, which contains two components : toner and carrier. The mixture quality valuation is made by different techniques, among which the most used seems to be that one based on specific charge q_x determination of toner powders. In this case for measuring Q_x, toner is separated from mixture.

3.1. Description of the Proposed Device

The device general diagram is shown in Fig. 4, where CF_2 is the F a r a d a y cup, P_+ and P_- are the high voltage separation plates, S_1, ..., S_4—swit-

ches on the front panel, IN_2 and IN_3 —instruments for plates voltage control, [KV], BA — supplying block, SIT — high voltage supply, VE — electrovalve for compressed air, other inscriptions being the same as before.

3.2. Q_x Measurement

a) *Device Preparation.* First of all CF_2 and support SE must be cleaned (for a good insulation of CF_2). Then it is necessary to control high tension supply SIT and electrovalve VE to be in order and to set the compressed air pressure at about 4 Kgf/cm².

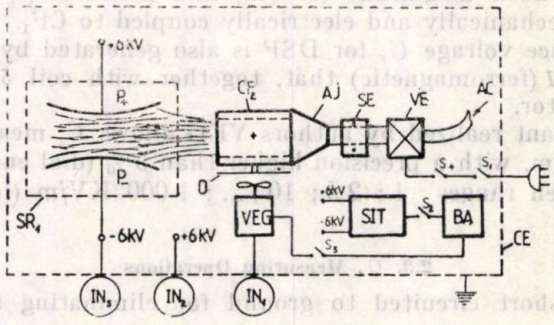

Fig. 4

b) *Measuring Operations.* 1. The powder to be measured (toner and carrier) is put in CF_2 which is weighed, screwed in the insulating nozzle AJ and then is discharged to ground (a contact is made between CF_2 and ground).

2. The lid of the apparatus is closed and SIT is driven (by acting S_2). After that VEG is driven by S_3 and IN_1 shows the spurious electric field E_p, if there is any.

3. S_4 is shortly pressed and IN_1 indicates E_x; thus Q_x can be calculated

$$(7) \qquad Q_x = k_2(E_x - E_p), \qquad Q_x \simeq k_2 E_x, \qquad (E_p \ll E_x),$$

where k_2 is the constant of the measuring device (CF_2—VEG) which can be: determined in the same way as k_1.

The explanation of E_x phenomenon is as follows; before air blowing the powder contained by CF_2 is electrically neutral (toner charge is equal to carrier charge). After blowing, toner leaves the cage, which, consequently, remains charged with Q_x corresponding to mass (m) of the toner expelled (carrier is retained by screen).

4. SIT is disconnected and CF_2 is again weighed. The difference between the two weighings allows mass m of expelled toner calculation, to which charge Q_x is attributed.

5. Using relation (2) the unknown q_x can be calculated.

The precision of q_x measurement using this method is reduced especially because of uncertainty in defining m (experiences show that not the entire mass of the expelled toner is charged). The method is however easy and quick (measurements are made without taking into account the phenomena present

in plates P compartiment) and precision is satisfactory for usual practice situations.

For more precise measurements it is necessary to use electric field separation.

c) *Determination of q_x by Electric Field Separation.* 1. The plates compartiment is prepared for use as follows : a) weighed sheets of paper are put on the bottom of the drawers SR_1 and SR_2 ; b) weighed aluminium foils are put on high voltage plates P (this can be done only if the mains swich is OFF and plates P are grounded, for eliminating electrocution dangers).

2. Operations 1, 2, 3 from point b) are made.

3. SIT being switched ON, drawer SR_1 with neutral powder is taken out easily,

4. SIT is disconnected (S_2) ; the potential of plates P disappears and some powder falls down in SR_2. Then plates P are easily shaken and aluminium foils together with the paper sheet from SR_2 are weighed. Finally, from the two weighings mass m is determined from relation (2).

Observations. 1. If the powder separated by plates P is not carefully weighed it is possible the error in determining q_x to be as great as in b) technique (by weighing CF_2).

2. If it is necessary to know the amounts of powder positively or negatively charged, a separating line on SR_2 axis may be mounted.

3. If it is necessary to know the amount of neutral powder, the deposited powder in SR_1 is weighed.

4. Conclusions

1. The proposed apparatus allows to determine the specific charge (q_x) of unmixed powders, between 10^{-8} ... 10^{-11} C/G, with a 5 ... 7% accuracy.

2. The set also makes possible q_x measurement for mixtures of powders, with or without high voltage electric field separation, with a precision higher than 10 ... 15%.

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APARAT PENTRU MĂSURAREA ELECTRIZĂRII TRIBOELECTRICE A PULBERILOR IZOLANTE

(Rezumat)

Se prezintă un aparat ce permite măsurarea sarcinii triboelectrice la pulberi, cu sau fără separare în cîmp electric de înaltă tensiune (6 kV) ; limite de măsură : 10^{-7} ... 10^{-10} C.

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D.C. 621.316

ELECTRONICALLY COMPENSATED INDUCTIVE CURRENT DIVIDER

BY

ELENA CREŢU, M. LUCANU, ADRIAN CREŢU and CARMEN FILIPIUC

1. Introduction

The inductive current divider (ICD), used when the galvanic insulation between the primary and the secondary circuits is not compulsory, can assure superior measuring accuracy in comparison to the typical one of a current transformer.

So, from the expression of the complex error $\underline{\Delta}$ of the noncompensated ICD (fig. 1) [6]:

$$\underline{\Delta} = \frac{\underline{Z}_2 - \underline{Z}_1(N_2 - N_1)/N_1}{(N_2/N_1)^2 \underline{Z}_0} + \frac{\underline{Z}_s}{(N_2/N_1)^2 \underline{Z}_0} = \underline{\Delta}_i + \underline{\Delta}_s, \tag{1}$$

where : $\underline{Z}_1$ and $\underline{Z}_2$ represent the impedances of the windings having N_1 and, respectively, $N_2 - N_1$ turns, $\underline{Z}_s$ — the load impedance, and $\underline{Z}_0$ — the no load operation impedance of the ICD, it has been found that with the condition :

$$\frac{\underline{Z}_1}{\underline{Z}_2} = \frac{N_1}{N_2 - N_1}, \tag{2}$$

which is possible to be achieved by a careful construction of the ICD winding [7], the component:

$$\underline{\Delta}_i = \frac{\underline{Z}_2 - \underline{Z}_1(N_2 - N_1)/N_1}{(N_2/N_1)^2 \underline{Z}_0} \simeq 0, \tag{3}$$

owed to the internal load, can be in actual fact eliminated.

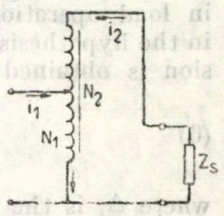

Fig. 1

Consequently, the measuring accuracy of the ICD is practically limitated only by the errors owed to its load impedance $\underline{Z}_s$ (the $\underline{\Delta}_s$ component from (1)), in the same way as in case of the two-stage current transformers [3], [6], but obviously, the ICD having a smaller price.

In the present paper it is studied the possibility to compensate automatically the errors of the ICD, by application of some electronic methods used in case of current transformes [1], [2], [8], ..., [11].

2. Methods for the Compensation of the ICD Errors by Means of Electronic Devices Dimensioned According to the Load Power

The electronic compensation of the complex error $\underline{\Delta}_s$, caused by the load $\underline{Z}_s$ of the ICD, can be achieved, in the same way as in case of the current transformer [1], [11], using the electronic amplifiers operating either as a

current converter (Fig. 2), respectively as a current—voltage converter (Fig. 4) or operating as a repeater (Fig. 3).

In case of compensation diagrams presented in Fig. 2, respectively in Fig. 3, the secondary winding N_2 of the ICD being practically short-circuit connected, the errors due to the load impedance $\underline{Z}_s$ are eliminated for a relative large range of the $\underline{Z}_s$ values.

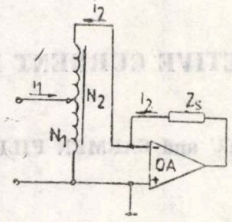

Fig. 2

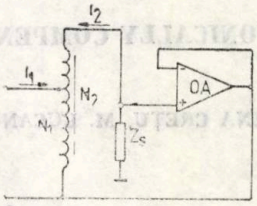

Fig. 3

So, on basis of equations written for the diagram in Fig. 2, which are :

$$\underline{I}_{10}=\underline{I}_1+\frac{N_2}{N_1}\underline{I}_2, \tag{4}$$

$$-\frac{N_2}{N_1}\underline{Z}_0\underline{I}_{10}\simeq(\underline{Z}_1+\underline{Z}_2)\underline{I}_2+\underline{Z}_1\underline{I}_1+\frac{\underline{Z}_S}{A}\underline{I}_2, \tag{5}$$

where $\underline{I}_1$, $\underline{I}_2$ represent the primary and the secondary currents of the ICD in load operation, and $\underline{I}_{10}$ is the primary current in noload operation, in the hypothesis that the condition (2) is achieved, the following final expression is obtained :

$$\underline{\Delta}_\alpha\simeq\underline{\Delta}_{s\alpha}=\frac{\underline{Z}_s}{(N_2/N_1)\underline{Z}_0A}=\frac{\underline{\Delta}_s}{A}, \tag{6}$$

where $\underline{\Delta}_s$ is the complex error of the non-compensated ICD, and A — the voltage amplification of the operational amplifier OA that has been used.

Similar results are also obtained for the compensation diagram presented in Fig. 3, where the equation of the ampere-turns has the form (4), and the voltage equation is :

$$-\frac{N_2}{N_1}\underline{Z}_0\underline{I}_{10}=(\underline{Z}_1+\underline{Z}_2)\underline{I}_2+\underline{Z}_1\underline{I}_1+\frac{\underline{Z}_s(1-A_{CM})}{A}\underline{I}_2, \tag{7}$$

where A_{CM} represents the common mode amplification of OA ($A_{CM}\ll A$).

From (4) and (7), condition (2) being respected, for the complex error of the electronically compensated ICD given in Fig. 3, the following expression is obtained :

$$\underline{\Delta}_\alpha\simeq\underline{\Delta}_{s\alpha}=\frac{\underline{Z}_s(1-A_{CM})}{(N_2/N_1)^2\underline{Z}_0A}=\frac{\underline{\Delta}_s(1-A_{CM})}{A}. \tag{8}$$

If the voltage amplification of the operational amplifier OA that has been used is large enough ($A \gg 1$), on basis of expressions (6) and (8) it results that, by application of compensation diagrams presented in Fig. 2, respectively in Fig. 3, an ICD with practically zero errors can be obtained (under 0.1 ppm) within a quite large range of values for the load impedance $\underline{Z}_s$ and for the working frequence.

In the compensation diagram proposed in Fig. 4, for the error compensation of an ICD, there is used a current—voltage converter [4], which, as well as in the previous cases, practically connects the secondary winding in short-circuit. This diagram can be used in the construction of that electronic measuring instruments to which the output signal u_e of the compensated ICD is requested to be only direct proportional to the secondary current i_2 (and not rigorously equal to it).

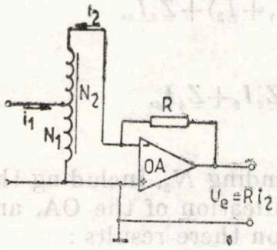

Fig. 4

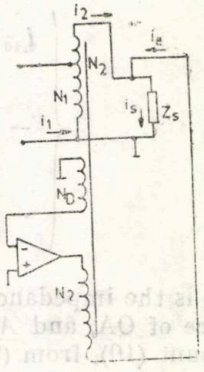

Fig. 5

In all the above studied compensation diagrams, the electronic device for compensation, OA, must deliver the whole power needed by the load $\underline{Z}_s$. Consequently, these diagrams are suitable for application only when the load current of the ICD has small values, the OA being of small power.

3. The Electronic Compensation of the ICD Errors by Means of Electronic Devices Dimensioned According to the Power Corresponding to the Error Current

The compensation of the ICD errors using an electronic device of small power (OA gives only the power corresponding to the error current), no matter how great the load power would be, can be achieved by means of the diagram presented in Fig. 5.

The annulation of the flux in the core of the ICD and, consequently, the annulation of its complex error $\underline{\Delta}$, is achieved by means of the demagnetizing ampere-turns $N_2\underline{I}_e$ produced by the compensation electronic device (operational amplifier OA and auxiliary winding N_2), which receives to its input the electromotive force induced in the detection winding N_D by the flux existent in the core of ICD.

The complex error $\underline{\Delta}_\alpha$ of the ICD compensated by this method, is defined by the ratio of the currents :

$$\frac{\underline{I}_s}{\underline{I}_1} = -\frac{N_1}{N_2}(1-\underline{\Delta}_\alpha), \tag{9}$$

where :

$$\underline{I}_s = \underline{I}_2 + \underline{I}_e \tag{10}$$

represents the current through the load impedance $\underline{Z}_s$, and $\underline{I}_e$ is the current delivered by OA in the auxiliary winding N_2.

Considering the equations of the diagram in Fig. 5 :

$$\begin{cases} \underline{I}_{10} = \underline{I}_1 - \dfrac{N_2}{N_1}\underline{I}_s, \\ -\dfrac{N_1}{N_1}\underline{Z}_0\underline{I}_{10} = \underline{Z}_2\underline{I}_2 + \underline{Z}_1(\underline{I}_1+\underline{I}_2) + \underline{Z}_s\underline{I}_s, \\ -\left(\underline{A}\dfrac{N_D}{N_2} + \dfrac{N_2}{N_1}\right)\underline{Z}_0\underline{I}_{10} = \underline{Z}_e\underline{I}_e + \underline{Z}_s\underline{I}_s, \end{cases} \tag{11}$$

where $\underline{Z}_e$ is the impedance of the auxiliary winding N_2, including the output impedance of OA, and $\underline{A}$ is the voltage amplification of the OA, and taking into account (10), from (9), after all calculation there results :

$$\underline{\Delta}_\alpha = \frac{\underline{Z}_2 - [(N_2-N_1)/N_1]\underline{Z}_1 + \underline{Z}_e + \underline{Z}_s(\underline{Z}_1+\underline{Z}_2)/\underline{Z}_e}{(N_2/N_1)^2\underline{Z}_0\{1+[\underline{Z}_1+\underline{Z}_2)/\underline{Z}_e](\underline{A}N_D/N_2+1]}. \tag{12}$$

If in the construction of ICD condition (2) is respected, and assuming that :

$$\frac{\underline{Z}_1+\underline{Z}_2}{\underline{Z}_e}\left(\underline{A}\frac{N_D}{N_2}+1\right) \gg 1, \tag{13}$$

expression (12) becomes simpler :

$$\underline{\Delta}_\alpha \simeq \frac{\underline{\Delta}_s}{\underline{A}\,N_D/N_2+1}\left(1+\frac{\underline{Z}_e}{\underline{Z}_1+\underline{Z}_e}\right), \tag{14}$$

where $\underline{\Delta}_s$ is the error due to the load impedance of the non-compensated ICD (s. Eq. (1)).

The reduction of error $\underline{\Delta}_\alpha$ to negligible values may be obtained by the use of an operational amplifier OA, having a great enough voltage amplification ($A \gg 1$).

In case of the analized diagram, the load current $\underline{I}_s \simeq \underline{I}_2$ is taken straight from the primary circuit of ICD (through the galvanic link), and the amplifier OA delivers only the power required for the compensation of the error $\underline{\Delta}_s$.

Another possibility to compensate electronically the errors of ICD, using small power electronic devices, is presented in Fig. 6. This scheme, derived from the one used in [9] for the errors compensation of a current transformer, can be applied to any ICD from outside, without any constructive alteration.

The values of resistances R_{N1} and R_{N2}, connected in the primary, respectively in the secondary circuit of the ICD, are calculated on basis of relation :

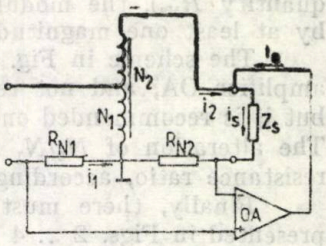

Fig. 6

$$\frac{R_{N2}}{R_{N1}} = \frac{I_1}{I_2} = \frac{N_2}{N_1}. \tag{15}$$

To the input of amplifier OA it is applied the voltage :

$$\underline{U}_i = R_{N1}\underline{I}_1 + R_{N2}\underline{I}_s = R_{N2}\frac{N_1}{N_2}\left(\underline{I}_1 + \frac{N_2}{N_1}\underline{I}_s\right) = R_{N2}\underline{\Delta}_\alpha, \tag{16}$$

which depends on the complex error $\underline{\Delta}_\alpha$ of ICD and from its output there is obtained the current :

$$\underline{I}_e = G\underline{U}_i, \tag{17}$$

where G represents the transfer conductance of the OA.

The current $\underline{I}_e$ modifies the current in the load impedance $\underline{Z}_s$ of the ICD from the value $\underline{I}_2$ to :

$$\underline{I}_s = \underline{I}_2 + \underline{I}_e. \tag{18}$$

The expression of the complex error $\underline{\Delta}_\alpha$ of the ICD compensated by means of the above analyzed method results from (9), on basis of the ampere-turns and the voltage equations of scheme in Fig. 6 :

$$\begin{cases} \underline{I}_{10} = \underline{I}_1 + \dfrac{N_2}{N_1}\underline{I}_2, \\ -\dfrac{N_2}{N_1}\underline{Z}_0\underline{I}_{10} = \underline{Z}_1\underline{I}_1 + (\underline{Z}_1 + \underline{Z}_2)\underline{I}_2 + (R_{N2} + \underline{Z}_s)\underline{I}_s \end{cases} \tag{19}$$

and taking into account (16), ..., (18).

After performing all calculation there is obtained :

$$\underline{\Delta}_\alpha = \frac{\underline{Z}_2 - [(N_2 - N_1)/N_1]\underline{Z}_1 + R_{N2} + \underline{Z}_s}{(N_2/N_1)^2\underline{Z}_0(1 - GR_{N2})}. \tag{20}$$

If condition (2) is respected at the construction of ICD, then expression (20) gives finally :

$$\underline{\Delta}_\alpha \simeq \frac{R_{N2} + \underline{Z}_s}{(N_2/N_1)^2\underline{Z}_0(1 - GR_{N2})}. \tag{21}$$

Comparing expressions (21) and (1) it is imediately found that, with the proposed method, the no-load impedance $\underline{Z}_0$ of the ICD increases $(1 - GR_{N2})$ times (but, at the same time, the load impedance $\underline{Z}_s$ increases with quantity R_{N2}), the modulus of the complex error getting slightly smaller, by at least one magnitude order [9].

The scheme in Fig. 6 presents the advantage of using a small power amplifier OA, and not request any constructive modification of the ICD, but it is recommended only for ICD with constant transformer ratio N_2/N_1. The alteration of N_2/N_1 ratio also requests the alteration of the R_{N2}/R_{N1} resistance ratio, according to (17).

Finally, there must be noticed that, in comparison with diagrams presented in Figs. 2 ... 4 which assure only the compensation of Δ_s errors determined by the load $\underline{Z}_s$ of the ICD, the diagrams presented in Figs. 5 and 6 have a compensation effect on both Δ_s and Δ_i components of the complex error of ICD.

4. Conclusions

1. The possibility to compensate automatically the errors of ICD by means of various electronic methods is studied.

2. There is demonstrated that when using the electronic amplifiers operating as a current or a current—voltage converter (Figs. 2 and 4), or operating as a repeater (Figs. 3), the errors of ICD can be reduced under 0.1 ppm, for a large range of values of the load impedance $\underline{Z}_s$ and of the working frequence, without any constructive change of this device. With these methods, the electronic amplifier must deliver the whole requested power by the load of ICD.

3. When the load current of ICD is of relatively great value, the schemes in Figs. 5 and 6 are recommended, where the electronic amplifier delivers only the power corresponding to te the current of the ICD. The diagram in Fig. 6 can be used without any constructive change of the ICD.

4. There is established that, unlike the current transformers, to which, usually, the electronic compensation of errors implies their constructive modification (the introduction of some auxiliary cores and windings [2], [8], etc.), in case of ICD, the compensation of errors can be made from the outside, simpler and cheaper.

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DIVIZOR INDUCTIV DE CURENT COMPENSAT ELECTRONIC

(Rezumat)

Se studiază posibilitatea compensării automate a erorilor divizorului inductiv de curent (DIC) prin diferite procedee electronice.

Se arată că, spre deosebire de transformatoarele de curent, la care compensarea electronică a erorilor presupune modificarea lor constructivă (introducerea unor miezuri și înfășurări auxiliare), la DIC compensarea erorilor poate fi efectuată „din exterior", mai simplu și la un preț de cost mai redus.

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D.C. 621.317 ; 621.34 ; 621.87

SUR LA DÉTERMINATION DE LA POSITION DU CHARIOT MOBILE DES GRUES À VOLÉE HORIZONTALE ET L'AFFICHAGE SYNOPTIQUE DU RÉSULTAT

PAR

M. CREŢU, C. SĂRMĂŞANU et LIVIU BRENIUC

1. Introduction

Le mouvement va-et-vient du chariot porte-crochet sur la volée horizontale d'une grue à tour est réalisé à l'aide d'un mecanisme moteur asynchrone—embrayage électromagnétique—réducteur mécanique qui entraine le tambour sur lequel les câbles de traction s'enroulent et se déroulent [3]. Pendant le fonctionnement l'affichage numérique du rayon de la grue devient insuffisant et il doit être doublé par une indication synoptique de la position instantanée du chariot ; celle-ci est plus suggestive et elle est mieux observée par l'opérateur.

2. Le procédé utilisé

Dans le schéma de principe de l'adapteur de mesure [1] le nombre retenu dans le compteur de longueur NL, qui en binaire représente la longueur du rayon de la grue, est ensuite décodé et affiché numériquement. La même information stockée dans NL est aussi utilisée pour l'affichage synoptique de la position du chariot porte-crochet dans la cabine de l'opérateur.

Le schéma électronique (Fig. 1) synthetisé pour l'affichage synoptique, utilise des circuits de décodage BCD/décimal et une barette de 25 diodes luminiscentes (LED) à l'aide desquelles on réalise une résolution de 2 m. Des 25 diodes luminescentes montées en ligne (Fig. 1) à un moment donné il n'y a qu'une seule qui s'allume, située sur la barette, conformément à la valeur de déplacement du chariot.

La valeur minimum du déplacement (2 m) saisie par l'affichage synoptique est suffisante dans l'exploitation ordinaire de la grue, parce que, maintenant, les jalons qui indiquent la position du chariot de charge sont montés de 10 m en 10 m sur la volée par rapport avec l'axe de la tour [3].

Aux entrées *A*, *B*, *C*, des décodeurs I-2, ..., I-6 (CDB 442) est couplé le groupe des sorties *B*, *C*, *D* du compteur de longueur NL qui stocke l'information sur la valeur du rayon en mètres ; aux entrées binaires *A*, *B*, *C*, du décodeur I-1 (CDB 442) est couplé le groupe des sorties *A*, *B*, *C* du même

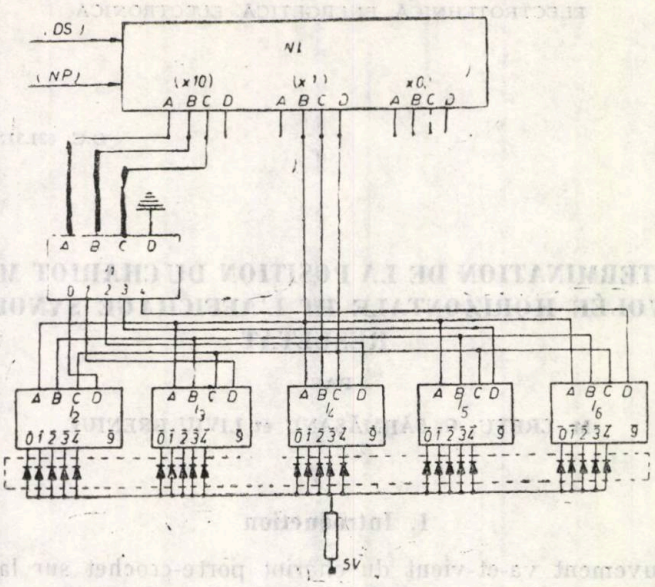

Fig. 1

compteur qui stocke à un moment donné la valeur de la longueur du rayon en groupes de dix mètres chacun (Tableau 1). L'entrée D du décodeur I-1 est couplée à la masse, les entrées D des décodeurs I-2, ..., I-6 sont couplées avec les sorties $\bar{0}$, ... $\bar{4}$, du décodeur I-1.

Tableau 1

No.	Valeur du rayon, [m]	0	10	20	30	40
1	A	0	1	0	1	0
2	B	0	0	1	1	0
3	C	0	0	0	0	1

Aux sorties $\bar{0}$, $\bar{1}$, $\bar{2}$, $\bar{3}$, $\bar{4}$, des décodeurs I-2, ..., I-6 sont couplés des groupes de 5 diodes luminescentes chacun qui, placées successivement sur la même ligne, forment la barette lumineuse.

On donne ci-dessous deux exemples numériques pour démontrer le fonctionnement du schéma. Soit, à un moment donné, le rayon de la grue egale à 36 m. Aux entrées A, B, C, du décodeur I-1, on applique le numéro :

A	B	C
1	1	0

du groupe des sorties ($\times 10$) du NL. La sortie 3 du même circuit est activée, elle devient 0 et par suite l'entrée D de l'intégré I-5 est aussi 0. Les autres entrées D des intégrés I-2, I-3, I-4 et I-6 non activées par le décodeur I-1, restent à l'état 1.

À l'entrée ($\times 1$) du compteur de longueur NL est inscrit le numéro 6, donc dans le code binaire on a :

A	B	C	D
0	1	1	0

La sortie A sur le degré ($\times 1$) de NL étant decouplée, le transfert de l'information vers les décodeurs est réalisé seulement par les sorties B, C, D, du même degré; parallèlement, à toutes les entrées des circuits I-2, ..., I-6 on applique le numéro :

A	B	C
1	1	0

Celui-ci actionne seulement à la sortie 3 du circuit I-5 qui a à l'entrée D la position 0. Pour les autres intégrés I-2, ..., I-6, puisque les entrées D sont situées dans l'état 1, on aura appliqué à l'entrée le numéro :

A	B	C	D
1	1	0	1

ou, en décimal, le numéro 11 qui dépasse la capacité des décodeurs et aucune diode des groupes de 5 ($\overline{0}$, ..., $\overline{4}$) couplées aux sorties des circuits I-2, ..., I-6 n'est pas activée. Par conséquent seulement la diode luminescente couplée à la sortie 3 s'allumera, ayant sur l'affichage synoptique une longueur de 34 ... 36 m.

Soit le rayon de la grue égale à 22 m. Aux antrées A, B, C, du décodeur I-1, on applique le numéro :

A	B	C
0	1	0

Du grupe des sorties ($\times 10$) du NL on active la sortie 2 du même circuit qui devient 0 et par conséquent l'entrée D du circuit I-4 est à l'état 0 aussi. Les autres entrées D de I-2, ..., I-6, pas activées par la décodeur I-1, restent à l'état 1.

À la sortie ($\times 1$) du NL est inscrit le numéro 2, respectivement, en binaire :

A	B	C	D
0	1	0	0

Puisque le transfert de l'information se réalise seulement par des sorties B, C, D, du même marché, sur toutes les entrées A, B, C des circuits I-2, ..., I-6 on applique le numéro :

A	B	C
1	0	0

qui rend active seulement l'entrée 1 de l'intégré I-4 qui à l'entrée D est à l'état 0. Aux autres intégrés, I-2, I-3, I-5, I-6, les entrées D étant à l'état 1, on aura à l'entrée le numéro :

A	B	C	D
1	0	0	1

ou, en décimal, le numéro 9 ; il faudrait donc qu'on active les diodes luminescentes $\overline{9}$, mais elles ne sont pas couplées sur la barette. Par conséquent les groupes LED ($\overline{0}$ ... $\overline{4}$) couplés aux sorties des décodeurs I-2, I-3, I-5 et I-6 restent éteints, seulement le LED couplé à la sortie 1 de l'intégré I-4, donc l'onzieme élément de la barette, s'allumera sur l'affichage synoptique et il représente une longueur de 20 ... 22 m.

Dans la cabine de l'opérateur la barette lumineuse est située à la base du graphique charge—rayon de tel façon que à une valeur, en mètres, du rayon

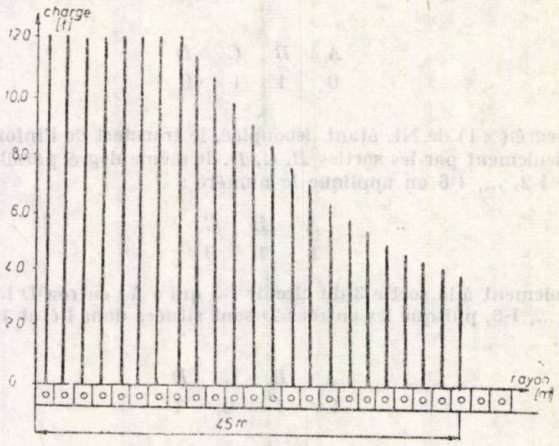

Fig. 2

(Fig. 2) indiquée synoptiquement par l'allumage d'un LED, correspond sur le graphique la charge maximum admise en tonnes.

3. Résultats expérimentaux et conclusions

Le schéma electronique (Fig. 1) réalisé pour l'affichage synoptique de la valeur du rayon a été testé en laboratoire sur un banc d'essai qui simule les conditions de travail de la grue. Le dispositif utilisé pour le teste (Fig. 3) contient : *1* — l'appareil à affichage synoptique réalisé ; *2, 3* — les éléments sensibles de proximité du type IPO5LPO24A ; *4* —la règle circulaire à tastes

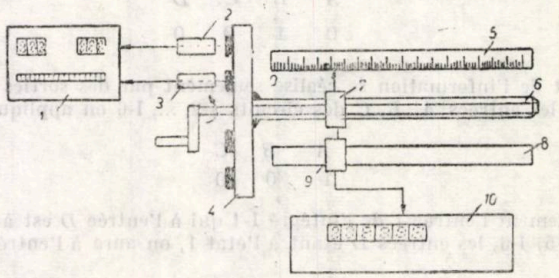

Fig. 3

de réluctance ; *5* — la règle métallique divisée en mm avec un longueur totale d'un mètre, du type I.E.P.A.M., Bîrlad ; *6* — le vis conducteur M 20×1.25 sur l'axe duquel la règle incrémentale à tastes de réluctance est fixée ; *7* — la règle incrémentale électro-optique du type Heindenhain LS 510 ; *8* — l'écrou

mobile solidarisé avec le curseur de la règle électro-optique ; *9* — la tête de lecture pour la règle Heindenhain ; *10* — l'instrument numérique de sortie du type Numerom-301 couplé avec la règle incrémentale.

La règle circulaire à tastes de réluctance a un pas [1], [2] angulaire de 40° et le rapport de conversion déplacement réel du chariot — déplacement du curseur règle Heindenhain est égal à 1 000, donc on peut considérer, que les valeurs des déplacements indiquées par l'instrument Numerom-301, après une multiplication à 10^3, représentent les déplacements réels du chariot.

Dans les limites de la résolution de l'affichage synoptique réalisé les résultats expérimentaux obtenus après le teste confirment l'exactitude de la solution proposée.

On peut adapter l'affichage synoptique de la position de moment du chariot de charge à n'importe quel type de grue à volée horizontale.

Dans la cabine de l'opérateur est indiqué d'une manière suggestive le rayon de la grue ; il peut aussi déterminer, à l'aide du graphique attaché à la barette lumineuse, la valeur de la charge maximum ; celle-ci ne doit pas être dépassée au crochet.

Sans modifications essentielles du schéma on peut réaliser des résolutions meilleures en augmentant le nombre des LED de la barette lumineuse.

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ASUPRA DETERMINĂRII POZIŢIEI CĂRUCIORULUI MOBIL AL MACARALELOR CU BRAŢ ORIZONTAL ŞI AFIŞAREA SINOPTICĂ A REZULTATULUI

(Rezumat)

Se prezintă schema electrică a afişării sinoptice a poziţiei căruciorului de sarcină a macaralelor cu braţ orizontal, care utilizînd 25 de LED-uri realizează o rezoluţie de 2 m la măsurarea razei ; concomitent se indică sugestiv şi valoarea sarcinii maxime admise la cîrlig. Aparatul realizat a fost testat pe un banc de probă în laborator.

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CONTROLLING DYNAMIC CHARACTERISTICS OF COMMUTATION ELECTRIC EQUIPMENT

BY

MARICEL ADAM and ADRIAN BARABOI

1. The continuous increase of electrical power consumption makes it necessary to use an ever larger number of high and medium voltage commutation electric equipment.

Proper operation of switching equipment influence the adequate exploitation of electrical plants, the continuity of electrical power supply and productiveness [1].

Determination of dynamic characteristics of switching equipment, reflecting to a good degree their operation, is done in keeping by means of the recording cylinder.

The present paper introduces a device based on a new method for the determination of the dynamic characteristic of commutation electric equipment.

2. Electrical performances of mechanical switching equipment are conditioned by providing some kinematic parameters like velocity and (total, in contact) stroke of mobile contact bar, these in their turn being conditioned by the driving gear of switching equipment.

With a view to converting the rotating motion given by the driving gear of the motion of translation on the level of the mobile contact bar, the crank and connecting rod assembly is frequently employed in the design of the electric commutation equipment.

By knowing the values of kinematic quantities of a point in the driving chain of the movement from the driving gear to mobile contact bars kinematic parameters of the mobile contact might be determined.

Taking into account the kinematics of the crank and connecting rod assembly with eccentricity [2] (Fig. 1), the angular stroke of the crank has been determined and divided into nine intervals delimited by ten marks (slits). By graphical or analytical methods the position of the connecting rod end (the A driving end of the mobile contact rod) can be successively determined for the ten marks.

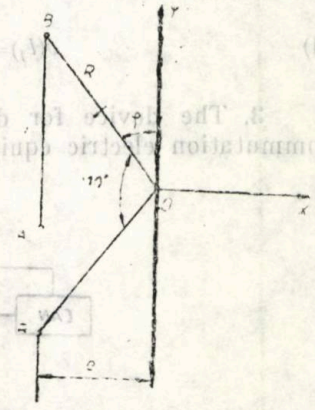

Fig. 1

Thus, depending on the φ angle made by the crank with its vertical passing through its shaft, the positions of the A connecting rod are given by the relation :

$$(1) \qquad Y_j = R\cos\varphi_j - L\sqrt{1-\left(\frac{R\sin\varphi_j - e}{L}\right)^2}, \quad (j=1, 2, \ldots, 9),$$

where : R — crank length, L — connecting rod length, e — eccentricity.

The designed device follows the determinations of angular stroke $\varphi(t)$ by measuring the time interval between two successive, marks thus between two successive positions of the connecting rod, on switching off or on the commutation equipment.

Duration Δt_j of the movement corresponding to the angle between the two successive marks, is determined by the relation :

$$(2) \qquad \Delta t_j = \frac{N_j}{f_T}, \quad (j=1, 2, \ldots, 9),$$

where : N_j represents the number of pulses recorded by the device for the respective field, f_T — the frequency of the signal generator.

The duration recorded from the beginning of the OFF (ON) stroke until reaching the mark k on the transducer sector is calculated by the relation :

$$(3) \qquad t_k = \sum_{j=1}^{k} \Delta t_j, \ (k=1, 2, \ldots, 9),$$

For determining the dynamic characteristic on turning OFF and ON the switch, curves $\varphi(t)$ are recorded, then by means of relation (1), space—time curves $y(t)$ are obtained. Velocity—time characteristics, $v(t)$, come out by differentiation, thus the velocity of the mobile contact corresponding to moment t_j can be considered of the form :

$$(4) \qquad v(t_j) = \left|\frac{Y_j - Y_{j-1}}{\Delta t_j}\right|, \quad (j=1, 2, \ldots, 9).$$

3. The device for determination of the dynamic characteristics of commutation electric equipment has the block diagram plotted in Fig. 2,

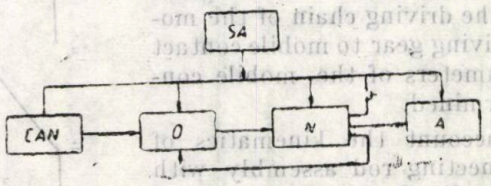

Fig. 2

and contains the following elements [3] : a) CAN — analog to digital converter ; b) D — distributor ; c) N — counter ; d) A — display unit ; e) SA — supply source.

The analog to digital converter includes two distinct parts, the transducer and the *CAN* itself. The transducer is made of a mechanical part directly linked on the kinematic chain of the switch and has got a dial under the form of a circle sector, on which radial marks are applied which will decompose the movement of the mobile contact into nine intervals. The electric part of the transducer is made a photodiode and a light source.

The pulses coming from the transducer will be taken over by *CAN*, and at its output ten rectangular pulses are obtained, of constant amplitude and different durations, corresponding to the passing of marks in front of the photodiode.

The pulses from the *CAN* are applied by means of the distributor on the ten counting channels of the counter.

The display unit is coupled for information reading on counting channels by means of a decodifier and of a selection circuit.

Conclusions

1. The device enables the checking up of electrical commutation equipment of medium and high voltage from the view point of driving gear operation, by storing and reproducing space—time characteristic of mobile contact both on switching OFF and ON of the tested equipment.

2. Time and average speed while OFF and ON can be calculated, and by graphically deriving the space—time characteristic, velocity—time characteristic can be obtained.

3. It is worth showing that by changing and adequate matching the transducer, the device can also be used in tracing the characteristics of some other processes.

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CONTROLUL PARAMETRILOR DINAMICI AI ECHIPAMENTELOR ELECTRICE DE COMUTAŢIE

(Rezumat)

Se prezintă o posibilitate practică de control a parametrilor mecanici ai echipamentelor de comutaţie, bazată pe cunoaşterea mărimilor care caracterizează dinamica unui punct din lanţul cinematic de transmisie a mişcării la echipajul mobil.

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SUR LE CALCUL OPTIMAL DES ÉLECTRO-AIMANTS À COURANT ALTERNATIF DE TYPE U-I MUNIS AVEC DES BAGUES DE DÉPHASAGE[1])

PAR

EUGEN HNATIUC

1. Introduction

L'analyse du fonctionnement des électro-aimants à courant alternatif de type U-I munis avec des bagues de déphasage afin d'éliminer les vibrations de leur armature mobile (Fig. 1), permet à écrire les expressions des flux magnétiques résultants $\underline{\Phi}_{e1}$ et $\underline{\Phi}_{e2}$ agissant dans la zone des pièces polaires [5] :

$$(1)\quad \begin{cases} \underline{\Phi}_{e1} = \dfrac{m}{m+1} \cdot \dfrac{1+\sin^2\varphi_{sc} - j\sin\varphi_{sc}\cos\varphi_{sc}}{1+3\sin^2\varphi_{sc}}\,\underline{\Phi}, \\ \underline{\Phi}_{e2} = \dfrac{1}{m+1} \cdot \dfrac{1+(2m+3)\sin^2\varphi_{sc} + jm\sin\varphi_{sc}\cos\varphi_{sc}}{1+3\sin^2\varphi_{sc}}\,\underline{\Phi}, \end{cases}$$

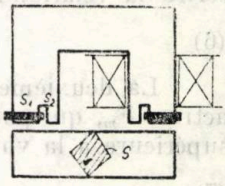

Fig. 1

où m est le rapport entre l'aire embrassée par la bague de déphasage (S_1) et l'aire qui n'est pas embrassée par cette bague (S_2), φ_{sc} — l'angle de déphasage de la spire en court-circuit ayant la résistance électrique R_{sc} et l'inductance L_{sc}, $\underline{\Phi}$ — le flux magnétique total du noyau de l'électro-aimant.

L'expression :

$$(2)\quad \Phi_{e2} = \frac{\Phi}{m+1}\sqrt{\frac{1+(m+1)(m+3)\sin^2\varphi_{sc}}{1+3\sin^2\varphi_{sc}}} \geqslant \frac{\Phi}{m+1} = \Phi_2,$$

confirme, ce qu'on savait déjà, la tendance de saturation magnétique dans cette partie du noyau.

En tenant compte des relations (1) et de l'angle de déphasage θ entre les flux magnétiques $\underline{\Phi}_{e1}$ et $\underline{\Phi}_{e2}$:

$$(3)\quad \cos\theta = \frac{\mathcal{R}e(\underline{\Phi}_{e1}\underline{\Phi}^*_{e2})}{\Phi_{e1}\Phi_{e2}} = \frac{1+(m+1)\sin^2\varphi_{sc}}{\sqrt{1+(m+1)(m+3)\sin^2\varphi_{sc}}},$$

on peut obtenir les expressions des composants moyen et oscillant de la force active agissant sur une pièce polaire du noyau magnétique :

$$(4)\quad \begin{cases} F_m = \dfrac{\Phi^2}{4\mu_0 S} \cdot \dfrac{1+(m+3)\sin^2\varphi_{sc}}{1+3\sin^2\varphi_{sc}} > \dfrac{\Phi^2}{4\mu_0 S}, \\ F_\sim = \dfrac{\Phi^2}{4\mu_0 S} \cdot \dfrac{\sqrt{1-2(m-3)\sin^2\varphi_{sc} + (m+1)(m+9)\sin^4\varphi_{sc}}}{1+3\sin^2\varphi_{sc}}. \end{cases}$$

1) Présentée au III-ième Symposium d'Appareils Électriques, Bucarest, 19—20 octobre 1989.

La relation (4_1) indique des valeurs de la force active moyenne plus grandes que celles obtenues en absence de la bague de déphasage.

Les relations (4) permettent aussi à évaluer la force active minimum pour une pièce polaire:

$$(5)\qquad F_{\min}=F_m-F_{\sim}=\frac{\Phi^2}{4\mu_0 S}\cdot\frac{1+(m+3)x-\sqrt{1-2(m-3)x+(m+1)(m+9)x^2}}{1+3x}\,,\ \sin^2\varphi_{sc}\underline{\Delta}x,$$

dont les valeurs supérieures aux celles de la force résistante assûrent l'élimination des vibrations de l'armature mobile de ces électroaimants.

2. Le calcul optimal des électro-aimants à courant alternatif de type U-I munis avec des bagues de déphasage

On peut utiliser les relations (4) et (5) pour le calcul optimal des électro-aimants mentionnés ci-dessus en tenant compte des conditions techniques concernant leur fonctionnement, à savoir: a) le fonctionnement normal, sans vibrations, est réalisé pour une tension d'alimentation du circuit de la bobine égale à $0{,}85\,U_n$; b) l'armature mobile reste dans la position finale, éventuellement avec des vibrations, pour une tension d'alimentation égale à $0{,}7\,U_n$.

La première condition met en évidence la valeur minimum de la force active, $F_{\min}$, qui pour la tension d'alimentation égale à $0{,}85\,U_n$ doit être supérieure à la valeur de la force résistante, F_r, de telle manière qu'on puisse écrire la relation :

$$(6)\qquad F_{\min}\big|_{0{,}85\,U_n}\geq F_r.$$

La deuxième relation met en évidence la valeur moyenne de la force active, F_m, qui pour la tension d'alimentation égale à $0{,}7\,U_n$ doit être aussi supérieure à la valeur de la force résistante, F_r, donc :

$$(7)\qquad F_m\big|_{0{,}7U_n}\geq F_r.$$

En tenant compte des expressions (2), (4) et (5), les relations (6) et (7) deviennent :

$$(8)\qquad \begin{cases} \dfrac{0{,}85^2B_{e2}^2S}{4\mu_0}\cdot\dfrac{1+(m+3)x-\sqrt{1-2(m-3)x+(m+1)(m+9)x^2}}{1+(m+1)(m+3)x}\geq F_r, \\ \dfrac{0{,}7^2B_{e2}^2S}{4\mu_0}\cdot\dfrac{1+(m+3)x}{1+(m+1)(m+3)x}\geq F_r, \quad B_{e2}=\dfrac{\Phi_{e2}}{S_2}, \end{cases}$$

où les valeurs maxima de l'induction magnétique B_{e2} sont d'habitude de 2 ... 2,2 T, (v. la relation (2)).

Les facteurs $0{,}85^2$ et $0{,}7^2$ des relations (8) suggèrent la dépendance linéaire des valeurs B_{e2} par rapport aux celles de la tension d'alimentation de la bobine, U, mais on peut améliorer la précision du calcul si on considère la dépendance non-linéaire $B(H)$ pour le matériel magnétique du noyau de l'électro-aimant.

En supposant connues les valeurs m et x, à l'aide des relations (8) on peut évaluer la valeur S de l'aire transversale du noyau magnétique de l'électro-aimant en choisissant la plus grande des valeurs obtenues à l'aide des realtions ;

$$(9)\quad \begin{cases} S^{*}=\dfrac{4\mu_0 F_r}{0{,}85^2 B_{\delta 2}^2}\cdot\dfrac{1+(m+1)(m+3)x}{1+(m+3)x-\sqrt{1-2(m-3)x+(m+1)(m+9)x^2}}, \\ S^{**}=\dfrac{4\mu_0 F_r}{0{,}7^2 B_{\delta 2}^2}\cdot\dfrac{1+(m+1)(m+3)x}{1+(m+3)x}. \end{cases}$$

Le calcul de ces électro-aimants est „optimal" si on obtient des valeurs minima de la section transversale, S.

À l'aide de représentation graphique des fonctions:

$$(10)\quad \begin{cases} g_1(x,\,m)=\dfrac{1}{0{,}85^2}\cdot\dfrac{1+(m+1)(m+3)x}{1+(m+3)x-\sqrt{1-2(m-3)x+(m+1)(m+9)x}^2}, \\ g_2(x,\,m)=\dfrac{1}{0{,}7^2}\cdot\dfrac{1+(m+1)(m+3)x}{1+(m+3)x}, \end{cases}$$

pour des valeurs imposées du rapport m (v. les relations (9) et la Fig. 2) on met en évidence un minimum pour la courbe $g_1(x)$ et des valeurs plus petites pour la courbe $g_2(x)$. On peut donc calculer la section transversale optimum du noyau magnétique de l'électro-aimant seulement à l'aide de la relation (9_1).

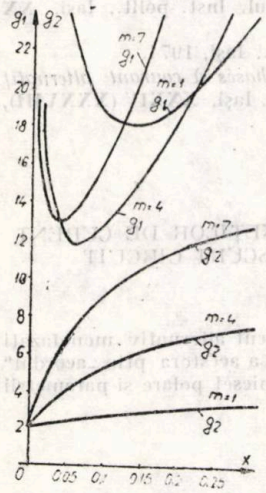

Fig. 2

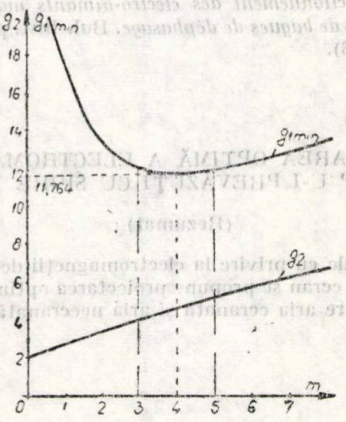

Fig. 3

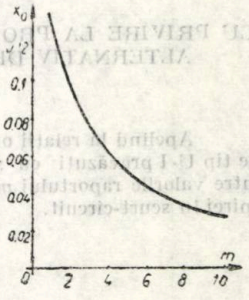

Fig. 4

C'est évident qu'on peut obtenir les plus petites valeurs S par la voie de „l'accord" des valeurs du rapport m et les valeurs x, c'est-à-dire pour $m=4$ et $x=0{,}0625$ (v. la Fig. 3).

La courbe $x_0(m)$, représentée dans la Fig. 4, indique aussi la nécessité de l'accord des valeurs m et x afin d'obtenir des valeurs minima de la fonction $g_1(x, m)$, qui ont évidemment une grande importance pour le calcul optimal de ces électro-aimants.

Il faut souliguer que les valeurs $x_0=\sin^2\varphi_{sc}$ seront considérées pour le calcul de la bague de déphasage.

3. Conclusions

Compte tenu des relations connues qui concernent le fonctionnement des électro-aimants monophasés de type U-I avec des bagues de déphasage [5] et aussi des conditions techniques correspondantes, on obtient des relations pour le calcul de la section transversale du noyau magnétique.

L'analyse de l'influence des valeurs m et x sur les valeurs de cette section transversale suggère la nécessité de „l'accord" de ces paramètres afin d'obtenir un calcul optimal.

On indique ainsi que les plus petites valeurs de la section transversale de ces électro-aimants seront obtenues pour $m=4$ et $x=0{,}0625$.

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Institut Polytechnique, Jassy
Chaire d'Electroénergétique

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CU PRIVIRE LA PROIECTAREA OPTIMĂ A ELECTROMAGNEŢILOR DE CURENT ALTERNATIV DE TIP U-I PREVĂZUŢI CU SPIRE ÎN SCURT CIRCUIT

(Rezumat)

Apelînd la relaţii originale cu privire la electromagneţii de curent alternativ monofazaţi de tip U-I prevăzuţi cu spire ecran se propune proiectarea optimală a acestora prin „acordul" între valorile raportului m dintre aria ecranată şi aria necranată a piesei polare şi parametrii spirei în scurt-circuit.

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ERROR-NUMBER INDICATING CODES IN ERROR CORRECTION (I)

BY

VALERIU MUNTEANU, M. LUCANU, ADRIANA PĂSTRĂVANU
and DANIELA TĂRNICERU

1. Definition Error-Number Indicating Codes

In what follows we consider that the code-word symbols as well as the received-word symbols belong to the Galois field GF(2).

The error number indicating codes provide the receiver the possibility to specify the number of errors that have occurred from the structure of the syndrome associated with the received word, calculated with the classical relation [5], [7]:

$$[Z]=[H][B]^T, \tag{1}$$

where $[H]$ is the parity-check matrix and $[B]^T$ is the transposed matrix of the received word:

$$[B]=[b_0\, b_1 \ldots b_{n-1}], \qquad b_i \in \mathrm{GF}(2), \qquad (i=\overline{1,\ n-1}). \tag{2}$$

Let us suppose that for a received word the calculated syndrome (by the help of (1)) indicates that e errors have occurred. Inverting an arbitrary symbol of the received word and calculating again the associated syndrome, two situations may arise: a) the new calculated syndrome indicates $e+1$ errors; b) the new calculated syndrome indicates $e-1$ errors.

In the first case it results evidently that the inverted symbol was correctly received, while in the second case, that it was not.

So, if one successively inverts the received-word symbols, one can correct all the e errors that have appeared, or can detect $e+1$ errors.

2. Relationships between Columns of the Parity-Check Matrix for e or Fewer Errors Correction

Let $[Z_1]$, $[Z_2]$, ..., $[Z_{e+1}]$ be the syndromes associated with the received words in which one, two or $e+1$ errors have respectively been introduced. Let also h_1, h_2, ..., h_n be the columns of the parity-check matrix.

If a single error occurs, the error word has the form:

$$[E_1]=[\ldots.e_i\ldots], \tag{3}$$

where $e_i=1$, all the other elements being zero for every $i=0, 1, ..., n-1$.

Taking (1) into account, when a single error appears, irrespective of its position, we can write :

$$[Z_1]=[h_1 h_2 \dots h_n][E_1]^T=[h_i]. \tag{4}$$

In order to be able to indicate a single error, irrespective of its position, it is necessary that $[Z_i]\neq[0]$, that is :

$$[h_i]\neq[0], \qquad (i=1, 2, ..., n). \tag{5}$$

When two independent errors occur, the error word is of the form :

$$[E_2]=[...e_k...e_j...], \quad k\neq j, \qquad (k, j=0, 1, ..., n-1), \tag{6}$$

where $e_k=e_j=1$, all the other elements being zero.

Taking (1) into account, when two independent errors occur we have

$$[Z_2]=[h_k]+[h_j], \tag{7}$$

where „+" denotes the addition in GF(2), used throughout this paper.

In order to be able to indicate all the combinations of two or fewer independent errors, it is necessary that :

$$[Z_2]\neq[0], \ [Z_2]\neq[Z_1] \text{ and } [Z_1]\neq[0]. \tag{8}$$

Taking (4) and (7) into account, relation (8) becomes

$$[h_i]+[h_j]+[h_k]\neq 0, \tag{9}$$

for every $i, j, k=1, 2, ..., n$. In other words, it means that in order to indicate two or fewer errors the parity-check matrix must be implemented so that the mod 2 sum of any three or fewer distinct colums be nonzero.

When three independent errors occur, the error word has the form :

$$[E_3]=[...e_m \dots e_p \dots e_q ...], \quad m\neq p\neq q, \ (m, p, q=0, 1, ..., n-1), \tag{10}$$

where $e_m=e_p=e_q=1$, all the other elements being zero.

The associated syndrome is then of the form :

$$[Z_3]=[h_m]+[h_p]+[h_q]. \tag{11}$$

In order to indicate three or fewer errors it is necessary that :

$$[Z_3]\neq[0], \ [Z_3]\neq[Z_2], \ [Z_3]\neq[Z_1], \tag{12}$$

equivalent with :

$$[h_j]+[h_k]+[h_m]+[h_p]+[h_q]\neq[0], \tag{13}$$

for every $j\neq k\neq m\neq p\neq q$ and $j, k, m, p, q=1, 2, ..., n$.

In other words, it means that in order to indicate three or fewer independent errors, it is necessary that the parity-check matrix must be implemented so that the mod 2 sum of five or fewer distinct columns be nonzero.

Following similar arguments it results

Theorem 1. *In order to indicate $e+1$ or fewer independent errors, the parity-check matrix must be so implemented that the mod 2 sum of any $2e+1$ or fewer distinct columns be nonzero.*

Let $GF(2^m)$ be the finite m-extension of $GF(2)$ and α a primitive element of $GF(2^m)$. It take place

Theorem 2. *The matrix*

$$[H]=\begin{vmatrix} 1 & 1 & 1 & \dots & 1 \\ 1 & \alpha & \alpha^2 & \dots & \alpha^{n-1} \\ 1 & \alpha^3 & (\alpha^3)^2 & \dots & (\alpha^3)^{n-1} \\ \cdot & \cdot & \cdot & \cdot & \cdot \\ 1 & \alpha^{2e-1} & (\alpha^{2e-1})^2 & \dots & (\alpha^{2e-1})^{n-1} \end{vmatrix} \tag{14}$$

enables the indication of $e+1$ *or fewer errors.*

Proof. In the matrix

$$[H^*]=\begin{vmatrix} 1 & \alpha & \alpha^2 & \dots & \alpha^{n-1} \\ 1 & \alpha^3 & (\alpha^3)^2 & \dots & (\alpha^3)^{n-1} \\ \cdot & \cdot & \cdot & \cdot & \cdot \\ 1 & \alpha^{2e-1} & (\alpha^{2e-1})^2 & \dots & (\alpha^{2e-1})^{n-1} \end{vmatrix} \tag{15}$$

the mod 2 sum of $r \leqslant 2e$ columns is nonzero [2], [3].

If one adds mod 2 any $2e+1$ distinct columns of the $[H]$ matrix, looking at its first row, it follows that an odd number of units are added and so, this sum is always nonzero.

3. First Method of Coding and Decoding Using Error-Number Indicating Codes

Let

$$[A]=[a_0\, a_1 \dots a_{n-1}],\ a_i \in GF(2), \tag{16}$$

be the transmitted code-word. One determines the check symbols of the code-word from the classical relation ([1], [5], [7]):

$$[H][A]^T=[0], \tag{17}$$

where $[H]$ is the parity-check matrix given by (14).

If the code-word is written in a polynomial form:

$$A(X)=a_0+a_1X+ \dots +a_{n-1}X^{n-1}, \tag{18}$$

then (17) is equivalent to:

$$A(\alpha^i)=0,\ (i=0, 1, 3, \dots, 2e-1). \tag{19}$$

In $GF(2^m)$, if (19) is true, in accordance with [1], [5], [7], it follows that:

$$A(\alpha^{2i})=0,\ (i=1, 2, \dots, e). \tag{20}$$

So, the relations (19) and (20) imply that:

$$A(\alpha^i)=0,\ (i=0, 1, 2, \dots, 2e). \tag{21}$$

Let $[B]$, given by (2), be the received word, or in a polynomial form:

$$B(X)=b_0+b_1X+ \dots +b_{n-1}X^{n-1},\ b_i\in \mathrm{GF}(2). \tag{22}$$

Supposing that the e errors in the received word are located on the positions $i_1, i_2, \dots, i_e$, the corresponding error word is

$$E(X)=E_{i_1}X^{i_1}+E_{i_2}X^{i_2}+ \dots +E_{i_e}X^{i_e}, \tag{23}$$

where $E_{i_1}=E_{i_2}= \dots =E_{i_e}=1$, if e errors have occurred in the received word and $E_{i_1}=E_{i_2}= \dots =E_{i_t}=1$, $E_{i_{t+1}}=E_{i_{t+2}}= \dots =E_{i_e}=0$ if $t<e$ errors have occurred.

The following relation takes place between the polynomials $B(X)$, $A(X)$ and $E(X)$:

$$B(X)=A(X)+E(X). \tag{24}$$

Substituting in (24) $X=\alpha^i$, $(i=0, 1, \dots, 2e)$, and taking (21) into account it follows that:

$$B(\alpha^i)=E(\alpha^i),\ (i=0, 1, \dots, 2e). \tag{25}$$

The notation can be simplified by denoting:

$$B(\alpha^i)=S_i \qquad \text{and} \qquad E_{i_k}\alpha^{i_k}=Y_kX_k. \tag{26}\ (27)$$

In the relation (26) the sums S_0 and S_{2i-1}, $(i=1, 2, \dots, e)$, can be calculated from the received word $B(X)$, successively substituting X by α^0, α^1, α^3, ..., α^{2e-1}. These sums can however be easier calculated from the remainder obtained by dividing the received word by $G(X)$, the generator polynomial of the code.

Indeed, if the code-words have the roots α^0, α^1, α^3, ..., α^{2e-1} and if we denote by $M_i(X)$ the minimal polynomials of the elements α^0, α^1, α^3, ..., α^{2e-1}, then:

$$G(X)=\mathrm{LCM}[M_0(X)\ M_1(X) \dots M_{2e-1}(X)], \tag{28}$$

where LCM denotes the least common multiplier.

On the other hand, we can write;

$$B(X)=G(X)Q(X)+R(X), \tag{29}$$

where $R(X)$ is the remainder of the division of the received word by $G(X)$, the generator polynomial of the code, and $Q(X)$ is the quotient. Substituting in (29) $X=\alpha^i$, $(i=0, 1, 3, \dots, 2e+1)$, and having in mind that $G(\alpha^i)=0$, it implies that:

$$B(\alpha^i)=R(\alpha^i),\ (i=0, 1, 3, \dots, 2e-1). \tag{30}$$

From (26) and (30) it follows that the sums S_i can be also calculated from the relations:

$$S_i=R(\alpha^i),\ (i=0, 1, 3, \dots, 2e-1). \tag{31}$$

The sums S_{2i}, $(i=1, 2, \dots, e)$, can be determined ([1], [5], [7]) with relation:

$$S_{2i}=S_i^2, \quad (i=1, 3, \ldots, e). \tag{32}$$

Since $X_k=\alpha^{i_k}$, where α is a primitive element in GF(2^m), it follows that $X_k\neq 0$, while some of the amounts Y_k can be zero, specifying that on the respective positions no errors have occurred.

From (23), (25),..., (27) it follows that :

$$S_i=\sum_{k=1}^{e} Y_k X_k^i. \tag{33}$$

It results

Theorem 3. *The rank of the matrix*

$$[M_s]=\begin{bmatrix} S_{e-1} & S_{e-2} & \ldots & S_1 & S_0 \\ S_e & S_{e-1} & \ldots & S_2 & S_1 \\ \ldots & \ldots & \ldots & \ldots & \ldots \\ S_{2e-2} & S_{2e-3} & \ldots & S_e & S_{e-1} \end{bmatrix} \tag{34}$$

indicates the number of errors that have occured in the received word.

Proof. Recalling the expressions for the sums S_i, given by the relation (33), the matrix M_e, given by (34), can equivalently be written as :

$$[M_e]=\begin{bmatrix} 1 & 1 & \ldots & 1 \\ X_1 & X_2 & \ldots & X_e \\ \ldots & \ldots & \ldots & \ldots \\ X_1^{e-1} & X_2^{e-1} & \ldots & X_e^{e-1} \end{bmatrix} \begin{bmatrix} Y_1X_1 & 0 & \ldots & 0 \\ 0 & Y_2X_2 & \ldots & 0 \\ \ldots & \ldots & \ldots & \ldots \\ 0 & 0 & \ldots & Y_eX_e \end{bmatrix} \begin{bmatrix} X_1^{e-2} & X_1^{e-3} & \ldots & 1 & X_1^{-1} \\ X_2^{e-2} & X_2^{e-3} & \ldots & 1 & X_2^{-1} \\ \ldots & \ldots & \ldots & \ldots & \ldots \\ X_e^{e-2} & X_e^{e-3} & \ldots & 1 & X_e^{-1} \end{bmatrix} \tag{35}$$

To simplify, we denote by

$$\det[M_e]=|M_e|. \tag{36}$$

It follows then :

$$\begin{bmatrix} Y_1X_1 & 0 & \ldots & 0 \\ 0 & Y_2X_2 & \ldots & 0 \\ \ldots & \ldots & \ldots & \ldots \\ 0 & 0 & \ldots & Y_eX_e \end{bmatrix} = X_1X_2\ldots X_e \begin{bmatrix} Y_1 & 0 & \ldots & 0 \\ 0 & Y_2 & \ldots & 0 \\ \ldots & \ldots & \ldots & \ldots \\ 0 & 0 & \ldots & Y_e \end{bmatrix} \tag{37}$$

and

$$\begin{bmatrix} X_1^{e-2} & X_1^{e-3} & \ldots & 1 & X_1^{-1} \\ X_2^{e-2} & X_2^{e-3} & \ldots & 1 & X_2^{-1} \\ \ldots & \ldots & \ldots & \ldots & \ldots \\ X_e^{e-2} & X_e^{e-3} & \ldots & 1 & X_e^{-1} \end{bmatrix} = \frac{1}{X_1X_2\ldots X_e} \begin{bmatrix} X_1^{e-1} & X_1^{e-2} & \ldots & X_1 & 1 \\ X_2^{e-1} & X_2^{e-2} & \ldots & X_2 & 1 \\ \ldots & \ldots & \ldots & \ldots & \ldots \\ X_e^{e-1} & X_e^{e-2} & \ldots & X_e & 1 \end{bmatrix}. \tag{38}$$

Using (37) and (38), the determinant of the matrix $[M_e]$, given by (35), can be calculated as follows. The determinant M_e is nonzero if all the three

determinants from the right hand side of (39) are nonzero. Since the first and the third are V a n d e r m o n d e type determinant, they are nonzero when

$$(39)\quad [M_e]=\begin{bmatrix} 1 & 1 & \dots & 1 \\ X_1 & X_2 & \dots & X_e \\ \dots & \dots & \dots & \dots \\ X_1^{e-1} & X_2^{e-1} & \dots & X_e^{e-1} \end{bmatrix}\begin{bmatrix} Y_1 & 0 & \dots & 0 \\ 0 & Y_2 & \dots & 0 \\ \dots & \dots & \dots & \dots \\ 0 & 0 & \dots & Y_e \end{bmatrix}\begin{bmatrix} X_1^{e-1} & X_1^{e-2} & \dots & X_1 & 1 \\ X_2^{e-1} & X_2^{e-2} & \dots & X_2 & 1 \\ \dots & \dots & \dots & \dots & \dots \\ X_e^{e-1} & X_e^{e-2} & \dots & X_e & 1 \end{bmatrix},$$

$$(40)\qquad X_1 \neq X_2 \neq \dots \neq X_e,$$

that is, when the e errors that occur in transmission are located on different positions in the received word.

The second determinant in (39) is nonzero when :

$$(41)\qquad Y_1 \neq Y_2 \neq \dots \neq Y_e,$$

that is when e errors occur in the received word.

If in the received word $t<e$ errors have occurred, then a number of $e-t$ amounts Y_i are zero.

In order to effectively establish the number of the occurred errors in the received word, one proceeds as follows :

1. If $|M_e|=0$, then one row and one column are eliminated from the matrix $[M_e]$. For the sake of simplicity the last row and the first column are eliminated obtaining the matrix :

$$(42)\qquad [M_{e-1}]=\begin{bmatrix} S_{e-2} & S_{e-3} & \dots & S_1 & S_0 \\ S_{e-1} & S_{e-2} & \dots & S_2 & S_1 \\ \dots & \dots & \dots & \dots & \dots \\ S_{2e-1} & S_{2e-2} & \dots & S_{e-2} & S_{e-1} \end{bmatrix}.$$

If $|M_{e-1}|\neq 0$ it follows that $e-1$ errors have occurred.

2. If $|M_{e-1}|=0$, from the matrix (42) are eliminated the last row and the first column resulting the matrix :

$$(43)\qquad [M_{e-2}]=\begin{bmatrix} S_{e-3} & S_{e-4} & \dots & S_1 & S_0 \\ S_{e-2} & S_{e-3} & \dots & S_2 & S_1 \\ \dots & \dots & \dots & \dots & \dots \\ S_{2e-2} & S_{2e-3} & \dots & S_{e-1} & S_{e-2} \end{bmatrix}.$$

If $|M_{e-2}|\neq 0$ it follows that $e-2$ errors have occurred, and if $|M_{e-2}|=0$ it may be proceeded similarly, till the first nonzero determinant is found, whose order indicates the actual number of errors that have occurred in the received word.

The decoding procedure using error number indicating codes consists then of the following steps :

1. From the received word, $B(X)$, or from the remainder, $R(X)$, the sums S_i, $(i=0, 1, 3, ..., 2e-1)$, are calculated by the help of relation (26) or (31) respectively, and the sums S_{2i}, $(i=1, 2, ..., e)$, using (32).

2. The rank of the matrix $[M_e]$, given by (34), indicates the number of errors that have occurred in the received word. Let us suppose that this rank was determined to be $r \leqslant e$.

3. Every symbol of the received word is successively inverted. For instance, inverting b_j, $(j=0, 1, ..., n-1)$, the sums:

$$S_i^{(j)} = S_i + (\alpha^i)^j, \quad (i=0, 1, 2, ..., 2e-2), \tag{44}$$

are calculated.

4. Then, the S_i sums are substituted by $S_i^{(j)}$ in the matrix $[M_e]$ and in this way the matrix $[M_e^{(j)}]$ is obtained; if the rank of the matrix $[M_e^{(j)}]$ is $r+1$, that is $M_{r+1}^{(j)} \neq 0$, it results that the symbol b_j was correctly received and if the rank of this matrix becomes $r-1$, that is $M_{r-1}^{(j)} \neq 0$, then the symbol b_j was erroneously received.

4. Conclusions

On the basis of a new type of codes — error number indicating codes — a new method for error correction is presented.

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CODURI INDICATOARE ALE NUMĂRULUI DE ERORI FOLOSITE ÎN CORECȚIA ERORILOR (I)

(Rezumat)

Se definesc codurile indicatoare de erori. Se prezintă apoi o primă metodă de corecție a erorilor folosind aceste coduri.

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SPECTRAL WIDTH OF MSK TYPE SIGNALING WAVEFORMS USED IN QAM SYSTEMS

BY

PAUL COTAE and N.D. ALEXANDRU

1. Introduction

Among the constant envelope signals imposing themselves by their features, there are MSK signals (Minimum Shift Keying). Taking into account the error probability they have identical performances with those of PSK signals [2], ..., [4], [5].

The MSK signal can be considered [2], [5] as :

1° A particular case of a frequency-modulated signal with phase continuity (CPFSK), the signaling frequencies being $f_1=f_c-1/4T$ and $f_2=f_c+1/4T$.

The continuous character of the MSK signal is illustrated in Fig. 1, where we considered the data sequences $a_k(t)$ with the simbols $a_k=\pm 1$ and signaling interval T.

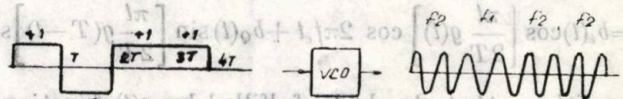

Fig. 1. — Continuous phase nature of MSK waveform.

In this case the MSK signal is described by the equation :

$$y(t)=\cos\left(2\pi f_c t+\frac{\pi a_k t}{2T}+\theta_k\right),\quad kT\leqslant t\leqslant(k+1)T, \tag{1}$$

$$s_1(t)=\cos(2\pi f_1 t+\theta_k),\quad s_{-1}(t)=\cos(2\pi f_2 t+\theta_k).$$

2° A particular case of the QPSK signal and it can be studied as a signal generated by a QAM modulation scheme [2] ... [4] :

$$y(t)=\cos\theta_k\cos\left(\frac{\pi t}{2T}\right)\cos 2\pi f_c t-a_k\cos\theta_k\sin 2\pi f_c t. \tag{2}$$

we must remark that the two functions :

$$\Phi_1(t)=\cos\frac{\pi t}{2T}\cos 2\pi f_c t,\quad \Phi_2(t)=\sin\frac{\pi t}{2T}\sin 2\pi f_c t \tag{3}$$

are orthogonal on an interval with the duration $2T$.

The MSK signal can be obtained with the arrangement given in Fig. 2. The following notation were used :

$$b_I(t)=g_I(t)a_I(t),\ b_Q(t)=g_Q(t)a_Q(t),$$
$$c(t)=b_I(t)\cos 2\pi f_c t,\ s(t)=b_Q(t)\sin 2\pi f_c t, \tag{4}$$

for the in-phase channel I and for quadrature channel Q. The resultant signal is :

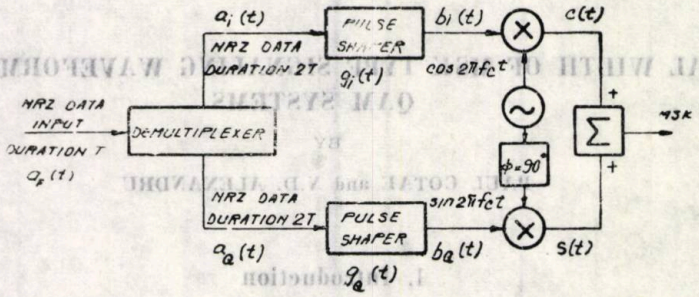

Fig. 2. — Method of modulating MSK.

$$y(t)=b_I(t)\cos\frac{\pi t}{2T}\cos 2\pi f_c t+b_Q(t)\sin\frac{\pi t}{2T}\sin 2\pi f_c t. \tag{5}$$

Taking this scheme as a startpoint, the MSK signal can be generalized, that is some other weighting circuits can be found leading to a general form signal :

$$y(t)=b_I(t)\cos\left[\frac{\pi t}{2T}g(t)\right]\cos 2\pi f_c t+b_Q(t)\sin\left[\frac{\pi t}{2T}g(T-t)\right]\sin 2\pi f_c t. \tag{6}$$

The conditions, that should be fulfilled by $g(t)$ function defined for the interval $[-T,\ T]$ and null for the rest, are [2] :

(7) 1. $g^2(t)+g^2(t-T)=A,\ 0\leqslant t\leqslant T,\ g^2(t)+g^2(t+T)=A,\ -T\leqslant t\leqslant 0,$

for maintaining the feature of constant envelope.

2. $g(-t)=g(t)$; $g(t)$ should exhibit even symetry around the mid-point of the interval $[-T,\ T]$, [2].

(8) 3. $\left(1-\frac{t}{T}\right)g(T-t)=1-\frac{t}{T}g(t)$;

$g(t)$ should have an odd symetry around the mid-point of the interval $[-T,\ 0]$, $[0,\ T]$.

(9) 4. $g^{(i-1)}(0)=0,\ (i=1,\ 2,\ \ldots,\ N-1).$

This ensures that the spectrum exhibits a fast spectral rolloff.

(10) 5. $g'(t)\leqslant 0\ \ 0\leqslant t\leqslant T,\ \ (t)\geqslant 0,\ -T\leqslant t\leqslant 0.$

Particular cases:

1°. $g(t)=0$, $\forall t\in(-T, T)$, $g(T)=1$. The OQPSK (offset or staggered QPSK) is obtained [2].

2°. $g(t)=1$, $\forall t\in[-T, T]$, the MSK signal is obtained.

3°. $g(t)=1-\dfrac{\sin 2\pi t/T}{2\pi t/T}$, $t\in[-T\ T]$ the SFSK signal (sinusoidal FSK) is obtained.

4°. $g(t)=1-\dfrac{\sum_{i=1}^{M} K_i \sin \alpha(2\pi t/T)}{\alpha(2\pi t)/M}$, $0\leqslant t\leqslant T$, where $K_i=\dfrac{(2i-2)}{2^{2i-2}[(i-1)!]^2(2i-1)}$ are the coefficients of the series expansion of arcsin x function. To be noticed that for $M=1$ and $\alpha=1$, 3° is obtained [5].

Some other particular cases of less interest are [3]:

5. $g(t)=(2t/T)^n$, $t\in[0, T]$.

6. $g(t)=\dfrac{1}{2}\left[\dfrac{1-\cos(2\pi t/T)}{t/T}\right]$.

2. Spectral Width of MSK Type Signaling Waveform Using Different Data Pulse Shape $g(t)$

If we consider that the data on the two channel are uncorrelated and the bits get equiprobables values, the MSK signal spectrum is the baseband one transposed around the carrier frequency (f_c).

Let $g(t)$ be defined on $[-T, T]$ fulfilling the conditions 1°...5°. We consider an equidistant division $\Delta=\{T_{-N}, T_{-N+1}, \ldots, T_{-1}, T_0, T_1, \ldots, T_{N-1}, T_N\}$ with the division norm $\|\Delta\|=T/N\to 0$, where $N\to\infty$.

Let the associated sum be:

$$g_\Delta(T)=\sum_{i=0}^{N} g(T-i)(T_{-i+1}-T_{-i})+\sum_{i=0}^{N} g(T_i)(T_{i+1}-T_i). \tag{11}$$

It comes out from (10) and (11) that:

$$g'_\Delta(t)=\frac{\mathrm{d}}{\mathrm{d}t}g_\Delta(t)=A_0[\delta(t+T)-\delta(t-T)]+\sum_{i=1}^{N} A_i[g(t-T_i)-g(t+T_i)], \tag{12}$$

where $A_i\geqslant 0$, $A_i=g(T_i)$.

The Fourier transform of the relation (12) is given by:

$$\mathrm{j}\omega G_\Delta(f)=A_0(\mathrm{e}^{\mathrm{j}2\pi fT}-\mathrm{e}^{-\mathrm{j}2\pi fT})+\sum_{i=1}^{N} A_i(\mathrm{e}^{\mathrm{j}2\pi fT_i}-\mathrm{e}^{-\mathrm{j}2\pi fT_i}) \tag{13}$$

or yet:

$$G_\Delta(f)=2T_0\frac{\sin 2\pi fT}{2\pi fT}+\sum_{i=1}^{N} 2T_iA_i\frac{\sin 2\pi fT_i}{2\pi fT_i}. \tag{14}$$

The spectrum of $G_\Delta(f)$ is different from that of $g(t)$. If $g(t)$ is sampled at a $1/T_s$ sampling rate, it is known that :

$$G_\Delta(f)=\left(T_s e^{-j2\pi fT_s}\frac{\sin \pi fT_s}{\pi fT_s}\right)\left[\frac{1}{T_s}\sum_{-\infty}^{+\infty}G\left(f-\frac{n}{T_s}\right)\right]. \tag{15}$$

But at the limit $T_s\to 0$, we get :

$$\lim_{T_s\to 0} G_\Delta(f)=G(f).$$

We are concerned with the appeareance of the first spectral null for $G_\Delta(f_e)$. In the particular case $A_i\neq 0$, $\forall \geq 1$, the rectangular waveform is obtained for $g(t)$, and the first null appears at $fT=1/2$. Excluding this case out of the question as if $T_i<T$, it results that for any other form of $g(t)$, $G_\Delta(f)$ will pass through zero for $fT>1/2$ values.

In the Table 1 these values were calculated for BPSK, QPSK, MSK, SFSK. In Fig. 3 the power spectral densities for OQPSK (PSK), MSK and SFSK are plotted. Analysing the relation (14) it is easily to see that more $A_i\neq 0$ coefficients we have, the bigger the null-to-null bandwidth will be.

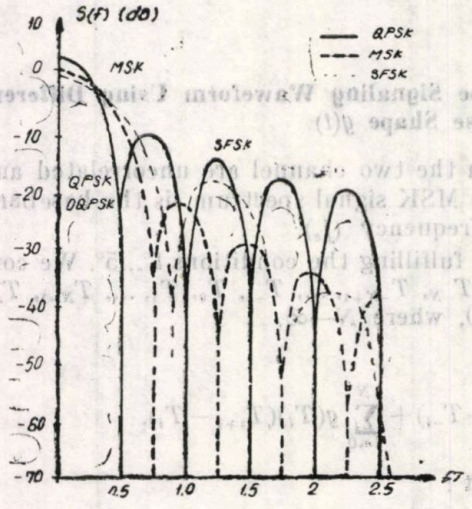

Fig. 3.— Power spectral density of QPSK, SFSK and MSK

Table 1

Modulation type	$2fT$
BPSK	2.00
QPSK	1.00
MSK	1.50
SFSK	1.72

3. Conclusions

The condition that should be fulilled by data pulse shape of the form cos $g(t)$ are given, particularising $g(t)$ for the cases met in the literature [1], ···,[6].

The spectrum of the MSK signal for any form of $g(t)$ is estimated and a comparison of the bandwidth occupied by the main lobe of the power spectral density for some other modulation types namely: BPSK, QPSK, MSK and SFSK is performed.

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LĂȚIMEA DE BANDĂ OCUPATĂ DE PRIMUL NUL SPECTRAL AL SEMNALELOR DE TIP MSK UTILIZATE ÎN SISTEMELE QAM

(Rezumat)

Se compară lățimea de bandă ocupată de lobul principal al densității spectrale de putere pentru diferite tehnici de modulație : BPSK, QASK, MSK și SFSK. Se demonstrează că alura căzătoare a densității spectrale de putere conduce la creșterea lățimii de bandă ocupată de primul nul spectral.

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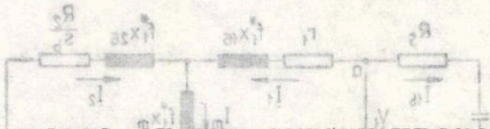


MATHEMATICAL MODEL OF THE SUBSYNCHRONOUS CONDENSERS BRAKING OF THE ASYNCHRONOUS THREE-PHASE MOTOR

BY

MIRCEA-SORIN OPRIŞOR and SIMONA-ANA CABA

1. Introduction

The connecting of a condensers battery to the terminals of the asynchronous three-phase motor has various effects, functions of the connection diagram, the values of the circuits parameters and the speed. If the primary feeding is cut off and the condensers battery is connected to the stator terminals while the rotational speed is high enough, the machine works as a self-excited generator through the agency of the remanent magnetism and of the condensers. The excitation frequency of the stator is the proper oscillation frequency of the circuit from the equivalent diagram represented in Fig. 1, and this depends on the rotation speed of the shaft n:

$$f_1 = \frac{pn}{1+s_b}, \ [s^{-1}], \tag{1}$$

and does not vary with the torque on the shaft [1]. For a high speed range, this frequency is smaller than the nominal one of the machine :

$$f_{1N} = \frac{pn}{1-s}.$$

We can conclude that the condensers braking is analogous to the adjustment of the asynchronous motor speed by the variation of the supply voltage frequency.

The torque produced by the motor can be used for different working machines, both for the rapid stopping (stop braking) [2], and for the limiting of the speed in under- and super-synchronous domains (maintaining braking) [3]. The main advantages of this braking are : high torque, rigidity of the mechanical characteristics, dynamic stability, simple adjustment [4], noiselessness, economic efficiency.

Among the possible solutions, this paper deals with the stop braking, especially in the reduced speed domain.

The analysis is based on the usual hypothesis for the asynchronous three-phase machine, namely : the symmetry of the machine and of the circuits connected to its terminals, the absence of the iron losses and the neglect of the saturation, of the discharge resistance effects of the condensers and of their inner resistance, of the connections resistance and of the diodes in conduction.

2. Operating Equations

Taking into account only the right-side of the stator terminals $a-o$ of the equivalent diagram (Fig. 1), the equations of the stator and rotor voltages have the following form [5]:

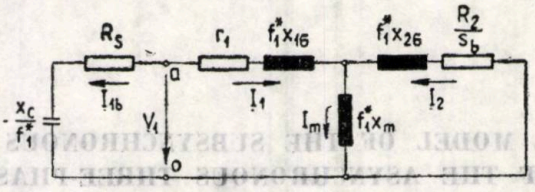


Fig. 1

$$(2)\quad \underline{V}_1=[r_1+\mathrm{j}f_1^*(x_{1\sigma}+x_m)]\underline{I}_1+\mathrm{j}f_1^* x_m\underline{I}_2,\quad 0=\mathrm{j}f_1^* x_m\underline{I}_1+\left[\frac{r_2}{s_b}+\mathrm{j}f_1^*(x_{2\sigma}+x_m)\right]\underline{I}_2,$$

where 1 is for stator, 2 for rotor, σ for straying, b for braking, m for magnetizing, the secondary parameters have the magnitudes values referred to the primary winding,

$$f_1^*=\frac{f_1}{f_{1N}}\quad\text{and}\quad s_b=\frac{s}{f_1^*}.$$

With the notations:

$$X_1=x_{1\sigma}+x_m,\quad X_2=x_{2\sigma}+x_m,\quad R_2=r_2+R,$$

the primary and the secondary currents may be obtained as:

$$(3)\quad \begin{cases} \underline{I}_1=\underline{V}_1\dfrac{R_2+\mathrm{j}s_b f_1^* X_2}{(r_1+\mathrm{j}f_1^* X_1)(R_2+\mathrm{j}f_1^* s_b X_2)+s_b f_1^{*2} x_m^2}, \\ \underline{I}_2=\underline{V}_1\dfrac{-\mathrm{j}f_1^* s_b x_m}{(r_1+\mathrm{j}f_1^* X_1)(R_2+\mathrm{j}f_1^* s_b X_2)+s_b f_1^{*2} x_m^2}. \end{cases}$$

Computing the torque of the asynchronous machine in the braking regime:

$$T_b=\frac{m}{2\pi f_1}\mathcal{R}e(\mathrm{j}f_1^* x_m\underline{I}_1^*\underline{I}_2),$$

an expression analogous to the Kloss equation is obtained:

$$(4)\quad T_b=\frac{2T_{kb}(1+\gamma_b)}{s_{kb}/s_b+s_b/s_{kb}+2\gamma_b},$$

where, by similitude with the normal asynchronous running, the following notations are used:

$$(5)\quad \begin{cases} T_{kb}=\dfrac{3pV_1^2}{4\pi f_1}\cdot\dfrac{s_{kb}f_1^{*2}x_m^2}{R_2(r_1^2+f_1^{*2}X_1^2)+r_1s_{kb}f_1^{*2}x_m^2}, \\ \gamma_b=\dfrac{R_2}{s_{kb}}\cdot\dfrac{r_1x_m^2}{(r_1X_2)^2+f_1^{*2}(X_1X_2-x_m^2)^2}, \\ s_{kb}=\dfrac{R_2}{f_1^*}\cdot\sqrt{\dfrac{r_1^2+f_1^{*2}X_1^2}{(r_1X_2)^2+f_1^{*2}(X_1X_2-x_m^2)^2}}. \end{cases}$$

As it can be noticed, all the values from Eq. (4) depend on the reported stator frequency f_1^*, thus on the shaft speed, so that the mechanical characteristic of the asynchronous machine — although it keeps the generally known aspect — should be an exception from the form corresponding to a normal running.

For practical calculus it is proper to obtain an expression depending on the constants of the machine corresponding to the normal asynchronous running. These result from the Eqs. (5) for $f_1^*=1$. By reporting the homologous values from the two running situations, we obtain :

$$(6)\quad \begin{cases} T_{kb}=T_k\lambda^2\dfrac{1+\gamma}{\sqrt{\delta\varepsilon}+\gamma},\quad \gamma_b=\dfrac{\gamma}{\sqrt{\delta\varepsilon}}, \\ s_{kb}=\dfrac{s_k}{\rho_2}\sqrt{\dfrac{\delta}{\varepsilon}},\quad s_b=\dfrac{f_1^*n_{0N}-n}{f_1^*n_{0N}}=\dfrac{s}{f_1^*}. \end{cases}$$

In the system (6), we have introduced the notations :

$$(7a)\qquad \delta=\frac{1}{f_1^{*2}}\left[1-\frac{1-f_1^{*2}}{1+(r_1/X_1)^2}\right],\quad \varepsilon=\left[1-\frac{1-f_1^{*2}}{1+(r_1/\sigma X_1)^2}\right]$$

and :

$$(7\,b)\quad \begin{cases} \lambda=\dfrac{V_1}{V_{1N}}\ -\ \text{reported voltage on the phase ;} \\ \rho_2=\dfrac{r_2}{r_2+R}\ -\ \text{reported rotor resistance ;} \\ \sigma=1-\dfrac{x_m^2}{X_1X_2}\ -\ \text{total dispersion coefficient.} \end{cases}$$

Replacing the expressions from the system (6) into the Eq. (4), it results :

$$(8)\qquad T_b=\frac{2T_k\lambda^2(1+\gamma)}{\dfrac{\delta f_1^*}{\rho_2}\cdot\dfrac{s_k}{s}+\dfrac{\varepsilon\rho_2}{f_1^*}\cdot\dfrac{s}{s_k}+2\gamma},$$

which represents a generalized form of the K l o s s equation, corresponding to the adjustment of the stator frequency of the asynchronous machine, and of the condensers braking of the same machine, respectively. For the machine

with the short-circuited rotor and with the stator fed to its nominal parameters, it is imposed that :

$$p_2 = \lambda = f_1^* = \delta = \varepsilon = 1,$$

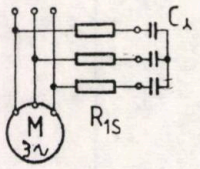
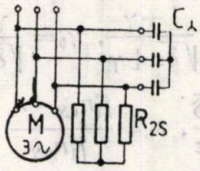
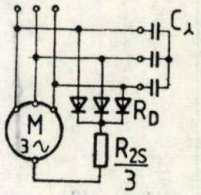
resulting, as a particular running case, the known form of the Kloss equation.

In Eq. (8) it is to notice that only the first two terms from the denominator (and sometimes λ from the numerator) depend on the shaft speed.

3. The Choosing of The Stator Diagram

The magnitude of the braking torque depends on the stator supply voltage and, implicitly, on the connection method of both the condensers (in star or delta) and of the stator supplementary resistances (s. Table 1). The

Table 1

a	b	c
		
$R_D = 0$ $R_{2S} = \infty$ $R_{3S} = 0$	$R_D = 0$ $R_{1S} = 0$ $R_{3S} = 0$	$R_D^+ = 0,\ R_D^- = \infty$ $R_{1S} = 0$ $R_{3S} = 0$
$\underline{V}_1 = \underline{I}_{1b}\left(R_{1S} - j\frac{x_C}{f_1^*}\right)$	$\underline{V}_1 = \underline{I}_{1b}\left[\frac{(x_C/f_1^*)^2}{R_{2S} - jx_C/f_1^*} - j\frac{x_C}{f_1^*}\right]$	$\underline{V}_1 = \underline{I}_{1b}\left[\frac{(x_C/f_1^*)^2}{(2R_{2S} - jx_C/f_1^*)}\right] - j\frac{x_S}{f_1^*}$

general equivalent diagram corresponding to a phase of the machine (left side of the terminals $a-o$—Fig. 2) can be used for the analysis. By writing the equations of the currents and voltages for these circuits [6] :

$$(9) \qquad \begin{cases} \underline{I}_{1b} = \underline{I}_C + \underline{I}_R = -\underline{I}_1, \\ \underline{V}_1 = R_{1S}\underline{I}_{1b} + (R_D + R_{2S})\underline{I}_R = R_{1S}\underline{I}_{1b} + (R_{3S} - jx_C/f_1^*)\underline{I}_C, \end{cases}$$

we obtain the general expression of the phase voltage :

$$(10) \qquad \underline{V}_1 = I_{1b}\left[R_{1S} + \frac{(R_D + R_{2S})(R_{3S} - jx_C/f_1^*)}{R_D + R_{2S} + (R_{3S} - jx_C/f_1^*)}\right].$$

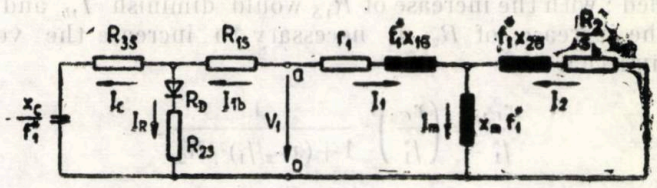


Fig. 2

The magnitude of the voltage V_1 depends on the magnitude of the supplementary resistances, of the reactance x_C and of the reported frequency f_1^*, as well as on the rotational speed of the shaft.

The voltage in the diagrams *a* and *b* (Table 1) is obtained from Eq. (10), specifying that $R_D=0$:

$$\underline{V}_1=\underline{I}_{1b}\left(R_{1S}+\frac{1}{\dfrac{1}{R_{2S}}+\dfrac{1}{R_{3S}-\mathrm{j}x_C/f_1^*}}\right). \tag{11}$$

The voltage in the case of the diagram *c* (Table 1) is:

$$\underline{V}_1=\frac{\underline{V}_{1R_D=0}+\underline{V}_{1R_D=\infty}}{2}=\underline{I}_{1b}\left[\left(R_{1S}+\frac{1}{\dfrac{1}{R_{2S}}+\dfrac{1}{R_{3S}-\mathrm{j}x_C/f_1^*}}\right)+\right.$$
$$\left.+\frac{1}{2}\left(R_{3S}-\mathrm{j}\frac{x_C}{f_1^*}-\frac{1}{\dfrac{1}{R_{2S}}+\dfrac{1}{R_{3S}-\mathrm{j}x_C/f_1^*}}\right)\right]. \tag{12}$$

By introducing the other particular conditions in the Eqs. (11) and (12), we get the voltages indicated in Table 1, for the most usual diagrams.

The influence of the stator diagram is to be noticed if we consider the voltages for the solutions *a* and *b*, the latter being written under the equivalent form:

$$\underline{V}_1=\underline{I}_{1b}\left[R_{2S}\frac{(x_{C2}/f_1^*)^2}{R_{2S}^2+(x_{C2}/f_1^*)^2}-\mathrm{j}\frac{x_{C2}}{f_1^*}\cdot\frac{R_{2S}^2}{R_{2S}^2+(x_{C2}/f_1^*)^2}\right]. \tag{13}$$

As δ, ε, γ, T_k and s_k do not depend on the parameters of the feed circuit, it results that for the same current I_{1b} (imposed by the shaft torque) and for the same reported frequency f_1^*, the torques will be equal when the real and the imaginary components of the voltages $\underline{V}_1$ are equal. Thus, it results that:

$$R_{2S}=\left(\frac{x_{C2}}{f_1^*}\right)^2\cdot\left[\frac{1}{2R_{1S}}\pm\sqrt{\left(\frac{1}{2R_{1S}}\right)^2-\left(\frac{f_1^*}{x_{C2}}\right)^2}\right], \tag{14}$$

whence the conditions

$$R_{2S}\geq\left(\frac{x_{C2}}{f_1^*}\right)\geq 2R_{1S} \tag{15}$$

must be fulfilled ; with the increase of R_{1S} would diminish I_{1b} and V_1, respectively, and the increase of R_{2S} is necessary to increase the voltage drop. Also it is found that :

$$\frac{x_{C1}}{f_1^*} = \left(\frac{x_{C2}}{f_1^*}\right) \cdot \frac{1}{1+(x_{C2}/f_1^*)^2/R_{2S}^2}. \tag{16}$$

Consequently :

$$x_{C2} > x_{C1}, \tag{17}$$

which represents a normal condition; as in the second case, the current I_{1b} passes through two parallel circuits.

By considering the voltages $\underline{V}_1$ for the diagrams *b* and *c* (Table 1), the former under the form shown in Eq. (13) and the latter under the equivalent form :

$$\underline{V}_1 = \underline{I}_{1b}\left[\frac{1}{2} R_{2S} \frac{(x_{C3}/f_1^*)^2}{R_{2S}^2+(x_{C3}/f_1^*)^2} - \mathrm{j}\frac{1}{2}\left(\frac{x_{C3}}{f_1^*}\right) \cdot \frac{2R_{2S}^2+(x_{C3}/f_1^*)^2}{R_{2S}^2+(x_{C3}/f_1^*)^2}\right], \tag{18}$$

by equalizing the real and the imaginary components — when there is an equality between the currents I_{1b}, the frequencies f_1^* and the resistances R_{2S} — we obtain :

$$\frac{x_{C3}}{f_1^*} = \left(\frac{x_{C2}}{f_1^*}\right)\sqrt{\frac{2}{1-(x_{C2}/f_1^* R_{2S})^2}}, \tag{19}$$

and, respectively,

$$R_{2S}^2\left(\frac{f_1^*}{x_{C2}}\right)^2 > 2R_{2S}^2\left(\frac{f_1^*}{x_{C3}}\right)^2 + 1. \tag{20}$$

Thus it results the condition :

$$x_{C3} > x_{C2} > x_{C1}. \tag{21}$$

The equality of the resistances values, when actually only $R_{2S}/3$ is used in the diagram *c* because of the compounding of the currents from the three phases, maintains the validity of the condition :

$$R_{2S} \geq 2R_{1S}. \tag{22}$$

From the conditions (21) and (22) we can conclude that the diagram *a* is the most favourable from the view point of the stator connection, as for the same values of R_S and x_C both the voltage V_1 and the torque T_b will be higher.

In all the analysed cases the supplementary resistances R_{2S} and the condensers have been star-connected, corresponding to the equivalent diagram from Fig. 2. From an economic view point it is convenient that, for these elements, we should choose the connection diagram which requires the smallest overall sizes and determines the minimum losses for the same braking torque. Thus, the star connection is to be preferred for resistances.

By means of the delta switching of the resistances R_{2S} from the diagram *b*, according to the condition (15), the solution becomes more disadvantageous when compared to the diagram *a*. The delta connection is preferable for condensers. In the case of high speed braking, the star connection could be used by obtaining a limitation of the phase voltage increase. Supplementary data are given in [7].

4. Conclusions

The asynchronous form of the mechanical characteristic corresponding to the condensers braking regime has been established.

The generalized form and the coefficients of the equation of the symmetrically three-phase asynchronous machine torque with variable frequency voltages have been determined. This is valid, as well, in the case of the condensers braking of this machine.

We have analysed different braking diagrams by means of condensers; thus, we have established the fact that the solution with stator supplementary series resistances is the optimal one from an economic point of view at reduced speeds. It has been also shown that the mounting of the delta condensers is to be preferred for the stop braking, while the star diagram is to preferred at the maintaining braking or as an intermediate braking.

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MODELUL MATEMATIC AL FRÎNĂRII SUBSINCRONE CU CONDENSATOARE A MOTORULUI ASINCRON TRIFAZAT

(Rezumat)

Se studiază pe cale analitică comportarea mașinii asincrone trifazate în regim de generator autoexcitat de la o baterie de condensatoare conectată în stator, folosită în scopul frînării electrice a sistemelor de acționare. În baza ecuațiilor stabilite a fost determinată forma generalizată a ecuației cuplului mașinii asincrone alimentată trifazat simetric cu tensiuni de frecvență variabilă și stabilită forma asincronă a caracteristicii mecanice ; de asemenea se face o comparație între diferitele scheme de conectare statorică. Se arată că, din punct de vedere economic, este favorabil ca rezistențele suplimentare statorice și condensatoarele să fie conectate după schema ce necesită gabaritele cele mai mici și determină, pentru același cuplu de frînare, pierderi minime.

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UNE NOUVELLE MÉTHODE D'IDENTIFICATION ET LOCALISATION D'UNE MISE À TERRE DANS UN RÉSEAU DE MOYENNE TENSION AYANT LE POINT NEUTRE ISOLÉ

PAR

JAN IGNAT

1. Considérations générales

Les méthodes utilisées pour l'identification et la localisation d'une mise a terre dans un réseau électrique de distribution de moyenne tension (RED—MT) ayant le point neutre isolé, permettent seulement à identifier le distributeur (D_i — Fig. 1) où s'est produit le défaut (le point K) sans offrir la possibilité de localisation ultérieure (après le débranchement du distributeur) du tronçon situé entre deux nodes (postes de transformation — PT) et des phases où s'est produit le défaut en vue de remédier les effets de celui-ci et de la remise en fonction du distributeur dans un intervalle de temps d'autant plus court que possible. Par conséquent la localisation du défaut, après le débranchement du distributeur, est effectuée dans un intervalle de temps de l'ordre des dizaines de minutes, par des recherches basées sur des opérations de commutation de proche en proche. Ce désavantage résulte de la nécessité de controler, en vue de l'identification du défaut, le déphasage entre les courants de deséquillibre (mesurés à l'aide d'un transformateur de courant de séquence homopolaire (TSH)) sur les départs du distributeur et la tension de déplacement du point neutre, $\underline{U}_{dd}$ (Fig. 1). Si, par exemple, c'est la phase R qui est mise à terre, alors le courant sur les départs sans défaut du distributeur est donné par [1] :

$$(1)\qquad \underline{I}^{s}_{\text{des},R} = \underline{I}_{RO_R} + \underline{I}_{SO_R} + \underline{I}_{TO_R} = \frac{-\mathrm{j}3\omega C_0 - \underline{U}f_{R\text{med}}}{1+\mathrm{j}3n\omega C_0 R_D},$$

tandis que le courant sur la phase où se produit le défaut est :

$$(2)\qquad \underline{I}^{d}_{\text{des},R} = \underline{I}_{pp_R} + (\underline{I}_{RO_R} + \underline{I}_{SO_R} + \underline{I}_{TO_R}) = (n-1)\frac{\mathrm{j}3\omega C_0 \underline{U}f_{R\text{med}}}{1+\mathrm{j}3n\omega C_0 R_D}.$$

Quant à la tension de déplacement du point neutre, U_{dd}, elle est accessible dans un seul point du RED—MT, à savoir la où est placé le transformateur qui crée le point neutre artificiel (TNA).

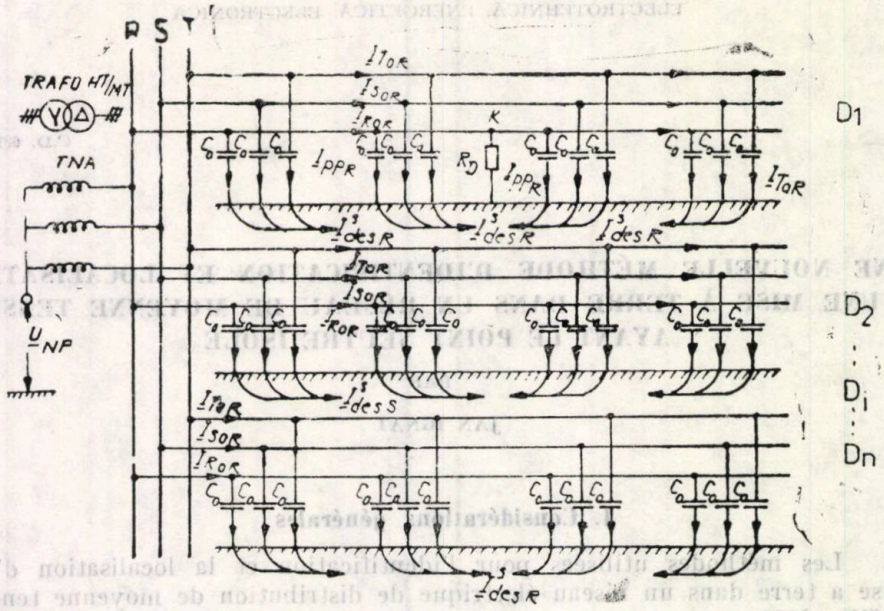

Fig. 1

Dans les relations (1) et (2) on a noté par n le nombre des distributeurs alimentés de la station de distribution, considérés de longueur égale, $\underline{U}_{f_{R\text{med}}}$ — la valeur moyenne de la tension sur la phase R de la source d'alimentation et R_0 — la résistance de mise à terre de la phase.

Selon la littérature de spécialité [1],...,[3], compte tenu que les courants $\underline{I}_{RO}$, $\underline{I}_{SO}$, $\underline{I}_{TO}$, ont, dans le cas d'une mise à terre, des valeurs sensiblement moindres que celles des courants en charge, on peut considérer que dans le régime de mise à terre monophasé les tensions de phase de la source et, respectivement, celles de ligne, ont les mêmes valeurs dans tout le réseau ; par conséquent on peut admettre que le composant de séquence inverse est pratiquement nul. On a noté les tensions de phase par $\underline{U}_{f_{R\text{med}}}$, $\underline{U}_{f_{S\text{med}}}$, $\underline{U}_{f_{T\text{med}}}$.

2. La méthode proposée

D'habitude les RED—MT ont la structure indiquée dans la Fig. 2, en étant alimentées par des transformateurs de force à connexions $Y_0\Delta$-5 et en alimentant, à leur tour, en cascade, par n distributeurs, des PT à transformateurs MT/0,4 KV, ayant la groupe de connexions ΔY-5 ou ΔY_0-11 [4].

Compte tenu des observation ci-dessus, les tensions de phase U_a, U_b, U_c sur la partie de basse tension de chaque PT (Fig. 2) restent patiquement inchangées dans un régime de mise à terre monophasée quelle que soit la valeur de la résistance R_0 et ont un déphasage strictement déterminé par rapport aux tensions du système triphasé sur la partie de MT du PT (Fig. 2) (qui coïncident avec les tensions du RED—MT). Il s'ensuit que pour chaque PT il existe un système triphasé de tensions de phase ($\underline{U}_a$, $\underline{U}_b$, $\underline{U}_c$), le même dans

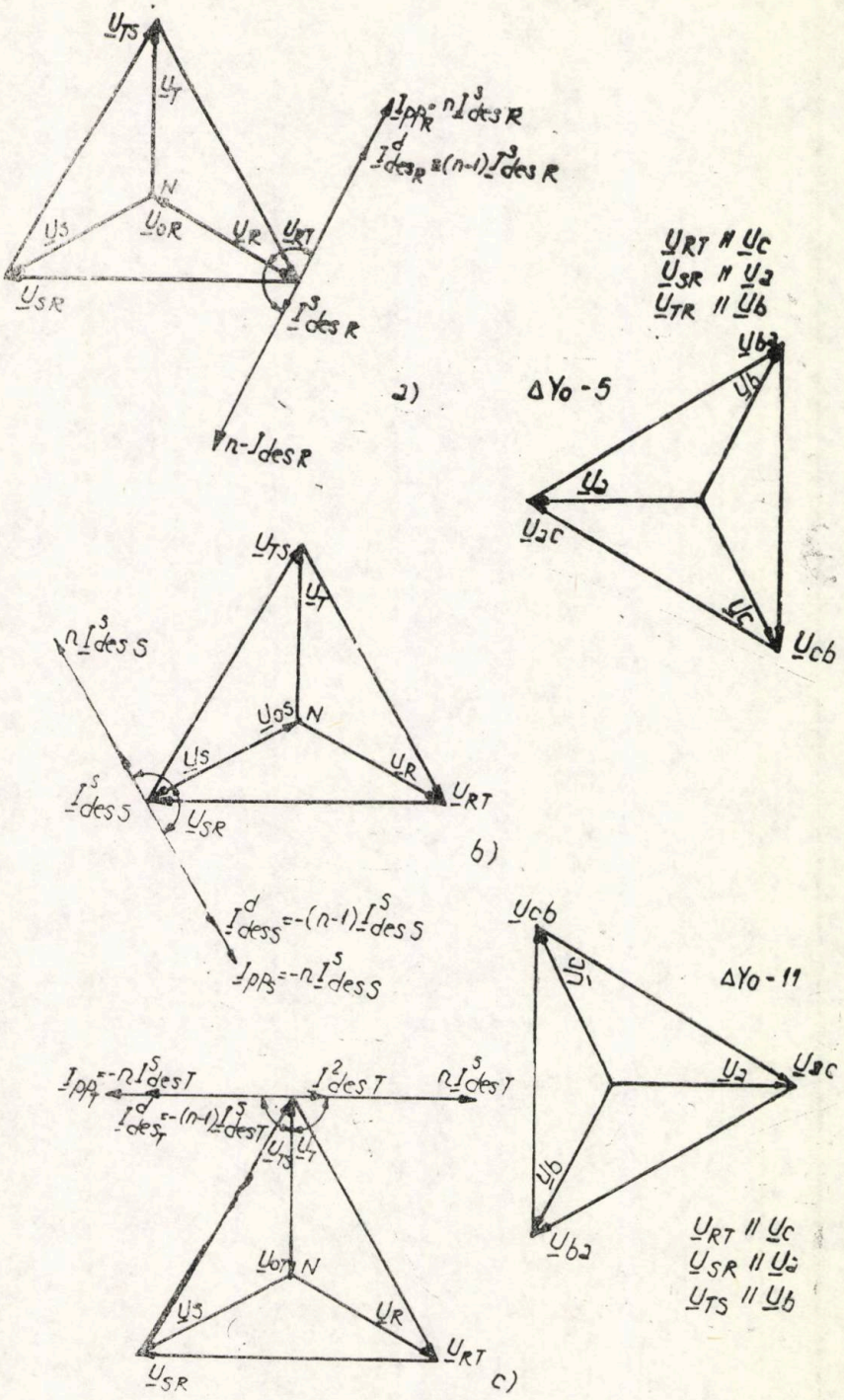

Fig. 3

tous les PT du RED—MT et par conséquent on peut l'adopter en tant que système de référence. Par surcroit il présente l'avantage d'être accessible directement, car il est bien connu le fait que dans les PT il n'y a pas, d'habitude, des transformateurs de mesure de tension (TT).

Il s'avère utile à déterminer les déphasages entre les courants de deséquilibre $\underline{I}^s_{des}$, $\underline{I}^d_{des}$, qui sont accessibles par l'intermédiaire des TSH (Fig. 2) et les tensions de phase $\underline{U}_a$, $\underline{U}_b$, $\underline{U}_c$, définies ci-dessus, dans l'hypothèse que

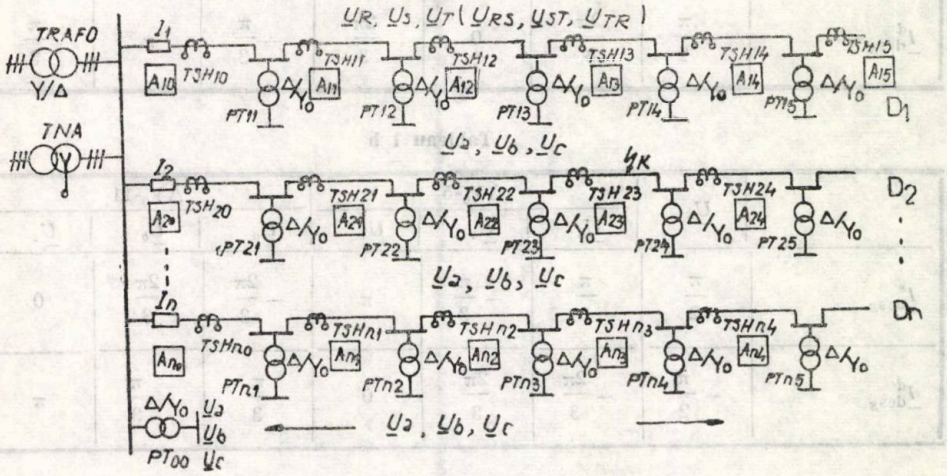

Fig. 2

la mise à terre est nette ($R_0=0$) pour les deux groupes de connexions ΔY_0-5(11). Dans ce but on a représenté, dans la Fig. 3, les diagrammes phasoriels des tensions de ligne $\underline{U}_{SR}$, $\underline{U}_{TS}$, $\underline{U}_{RT}$ et des celles de phase $\underline{U}_R$, $\underline{U}_S$, $\underline{U}_T$, dans les primaires des transformateurs de force du PT ; toutefois on a représenté les diagrammes phasoriels des tensions de ligne ($\underline{U}_{ba}$, $\underline{U}_{ca}$, $\underline{U}_{ac}$) et de phase ($\underline{U}_a$, $\underline{U}_b$, $\underline{U}_c$) dans les secondaires des mêmes transformateurs et des courants de deséquilibre (1), (2), dans le cas d'une mise à terre monophasée du RED—MT, succesivement pour la phase R (Fig. 3 *a*), S (Fig. 3 *b*) et T (Fig. 3 *c*).

Compte tenu des expression (1), (2) des courants de deséquilibre (qui sont de séquence homopolaire, $\underline{I}_{pp}=3\underline{I}_0$, déphasés avec $\pm\pi/2$ par rapport à la tension de la phase mise à terre) et des diagrammes phasoriels (Fig. 3), on a réuni, dans le Tableau 1, les valeurs des déphasages entre les courants $\underline{I}^s_{des}$, $\underline{I}^d_{des}$ et les tensions correspondantes du primaire et du secondaire du transformateur de force du PT, dans le cas d'une mise à terre succesive de la phase R (Tableau 1 *a*), S (Tableau 1 *b*) ou T (Tableau 1 *c*). Il résulte que les valeurs des déphasages entre les courants de deséquilibre $\underline{I}^s_{des}$, $\underline{I}^d_{des}$ d'une part et chaque tension de phase $\underline{U}_a$, $\underline{U}_b$, $\underline{U}_c$, d'autre part, sont différentes et uniquement déterminées par la phase mise à terre. D'autant plus, fonction de la groupe de connexions et de la phase mise à terre seulement un des angles de déphasage est nul.

De l'analyse effectuée ci-dessus il s'ensuit que par l'intermédiare du contrôle des déphasages entre les courants de deséquilibre $\underline{I}^s_{des}$, $\underline{I}^d_{des}$ et les

Tableau 1 a

	$\underline{U}_R$	ΔY_0-5			ΔY_0-11		
		$\underline{U}_a$	$\underline{U}_b$	$\underline{U}_c$	$\underline{U}_a$	$\underline{U}_b$	$\underline{U}_c$
$\underline{I}^s_{des_R}$	$\frac{\pi}{2}$	$-\frac{\pi}{3}$	π	$\frac{\pi}{3}$	$\frac{2\pi}{3}$	0	$-\frac{2\pi}{3}$
$\underline{I}^d_{des_R}$	$-\frac{\pi}{2}$	$\frac{2\pi}{3}$	0	$\frac{2\pi}{3}$	$-\frac{\pi}{3}$	π	$\frac{\pi}{3}$

Tableau 1 b

	$\underline{U}_S$	ΔY_0-5			ΔY_0-11		
		$\underline{U}_a$	$\underline{U}_b$	$\underline{U}_c$	$\underline{U}_a$	$\underline{U}_b$	$\underline{U}_c$
$\underline{I}^s_{des_S}$	$\frac{\pi}{2}$	$\frac{\pi}{3}$	$-\frac{\pi}{3}$	π	$-\frac{2\pi}{3}$	$\frac{2\pi}{3}$	0
$\underline{I}^d_{des_S}$	$-\frac{\pi}{2}$	$-\frac{2\pi}{3}$	$\frac{2\pi}{3}$	0	$\frac{\pi}{3}$	$-\frac{\pi}{3}$	π

Tableau 1 c

	$\underline{U}_T$	ΔY_0-5			ΔY_0-11		
		$\underline{U}_a$	$\underline{U}_b$	$\underline{U}_c$	$\underline{U}_a$	$\underline{U}_b$	$\underline{U}_c$
$\underline{I}^s_{des_T}$	$\frac{\pi}{2}$	π	$\frac{\pi}{3}$	$-\frac{\pi}{3}$	0	$-\frac{2\pi}{3}$	$\frac{2\pi}{3}$
$\underline{I}^d_{des_T}$	$-\frac{\pi}{2}$	0	$-\frac{2\pi}{3}$	$\frac{2\pi}{3}$	π	$\frac{\pi}{3}$	$-\frac{\pi}{3}$

tensions de phase $\underline{U}_a$, $\underline{U}_b$, $\underline{U}_c$ du réseau de basse tension (RBT) on peut localiser le distributeur, le tronçon situé entre deux PT du distributeur ainsi que la phase où s'est produit le défaut. Dans ces buts il s'avère nécessaire à utiliser, dans chaque PT, un TSH pour mesurer les courants de deséquilibre et un dispositif adéquat pour le contrôle des déphasages, compte tenu que les tensions $\underline{U}_a$, $\underline{U}_b$, $\underline{U}_c$, sont directement accessibles.

3. Dispositif de mesure

On présente, dans la Fig. 4, le schéma de principe du dispositif (A_{ij}) à l'aide duquel on peut controler les déphasages dans chaque PT du RED — MT (Fig. 2), selon la méthode proposée dans la Fig. 2. Celui-ci contient trois

comparateurs de phase (φ_R, φ_S, φ_T), un relais maximal de tension (U_p), un relais de temps (T), un élément de mémoire commandée (M_c) et une porte logique ET. Le relais maximal U_p assûre, d'une part, la desensibilisation du dispositif par rapport aux courants de deséquilibre ($\underline{I}_{des}$), déterminés par l'asymétrie du RED—MT, à savoir :

$$K_{sec}K_T I^s_{des} > U_p > K_{sec}K_p I_{des} \tag{3}$$

et d'autre part commande, au moment de l'apparition d'un courant de deséquilibre ($\underline{I}^s_{des}$, $\underline{I}^d_{des}$), déterminé par une mise à terre, l'élément de temps (T) qui, à son tour, après $\Delta t > t_{trans.} \approx 40$ ms, commande l'élément de mémoire (M_c) afin qu'il enregistre à sa sortie les valeurs des sorties des comparateurs (Q_1, Q_2, Q_3) dans ce moment là. Par cette voie on assure la desensibilisation par rapport au regime transitoire qui se produit lors d'une mise à terre [1], en permettant toutefois à lire des valeurs indiquées par les dispositifs même après le debranchement du distributeur en défaut. C'est dans ce but que la valeur donnée par Q_4 doit être retenue dans la mémoire.

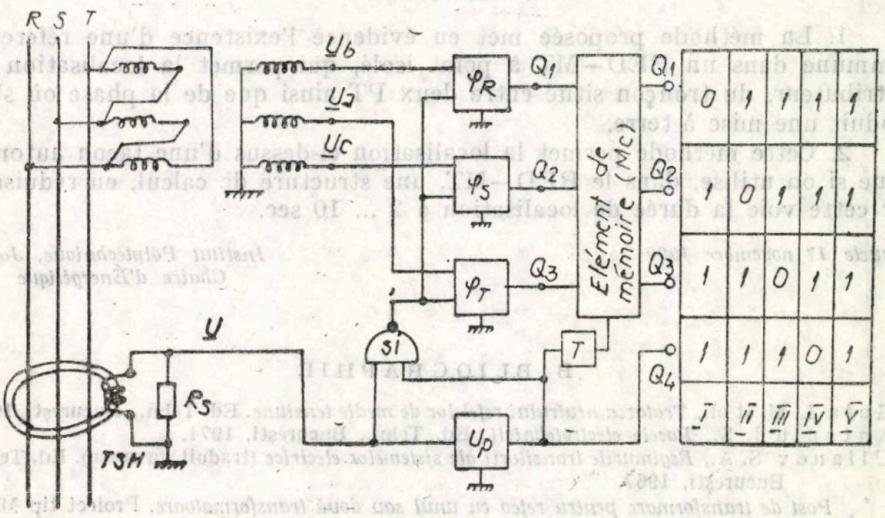

Fig. 4

Les comparateurs sont réalisés d'une telle façon de sorte qu'ils aient la grandeur de sortie située sur le niveau logique „0" lorsque les grandeurs d'entrée sont simphasiques où bien „1" lorsque une des grandeurs d'entrée est nulle. Cette dernière condition est assurée par la présence de la porte logique ET.

Les cinq états possibles des grandeurs de sortie Q_1, ..., Q_4 ont la signification suivante : I, II, III — mise à terre en aval sur la phase R, S, ou respectivement T ; IV — l'absence d'une mise à terre ; c'est l'état des dispositifs se trouvant sur le tracé du courant I^s_{des}, respectivement ceux situés sur les distributeurs sans défaut mais aussi ceux se trouvant en aval du point où se produit le défaut (K — Fig. 2).

Afin de localiser le distributeur et la phase mise à terre il s'avère suffisant à enregistrer l'état des dispositifs situés dans la station (A_{10} — Fig. 2) et puis celui des dispositifs placés sur le distributeur mis à terre (dont le dispositif A_{20} se trouve dans l'un des états I ... III) du côté d'amont vers aval jusqu'à l'identification du premier dispositif (A_{24}) se trouvant dans l'état V. Le tronçon à défaut est le premier du côté d'amont.

Il faut souligner que cette méthode permet à localiser le défaut même dans le cas ou le nombre de distributeurs est $n=2$; quant aux méthodes dont on dispose couramment elles sont applicables si seulement $n \geqslant 3$.

La méthode proposée a été vérifiée par voie expérimentale sur un modèle physique de RED—MT ; le dispositif dont le schéma de principe est représenté dans la Fig. 4, a constitué l'objet d'un brévet d'invention [6].

Évidemment si on dispose d'une structure automatique de calcul implementée dans le RED—MT [5], on peut effectuer d'une façon automate la lecture de l'état des dispositifs et par conséquent la durée de la localisation se reduit à 2 ... 10 sec.

4. Conclusions

1. La méthode proposée met en évidence l'existence d'une référence commune dans un RED—MT à point isolé, qui permet la localisation du distributeur, du tronçon situé entre deux PT ainsi que de la phase où s'est produit une mise à terre.

2. Cette méthode permet la localisation ci-dessus d'une façon automatique si on utilise, dans le RED—MT, une structure de calcul, en réduisant par cette voie la durée de localisation à 2 ... 10 sec.

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Chaire d'Énergétique

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METODĂ DE IDENTIFICARE ȘI LOCALIZARE A UNEI PUNERI LA PĂMÎNT ÎNTR-O REȚEA CU NEUTRUL IZOLAT

(Rezumat)

Se pune în evidență existența unei referințe comune și ușor accesibile în orice RED—MT cu neutrul izolat, care permite, prin controlul unor defazaje, localizarea distribuitorului, a tronsonului dintre două PT de pe distribuitor și a fazei pe care se produce o punere la pământ.

BULETINUL INSTITUTULUI POLITEHNIC DIN IAŞI
Tomul XXXVII (XLI) Fasc. 1—4, 1991
Secţia III
ELECTROTEHNICĂ. ENERGETICĂ. ELECTRONICĂ.

C.D. 621.311.153

SUR L'ÉVOLUTION DE LA CONSOMMATION DE L'ÉNERGIE ÉLECTRIQUE

PAR

GH. CÂRŢINĂ

1. Introduction

Grâce à son poids dans le bilan énergétique d'un pays, d'une part, et, d'autre part, aux changements structurels qui ont intervenu tout récemment, le secteur domestique constitue un sujet permanent pour les chercheurs du monde entier. Parmi les changements structurels les plus importants sont : l'augmentation de la surface habitable par logement, la diminution du nombre des personnes par logement, l'augmentation du nombre d'appareils électriques, l'utilisation de plus en plus répandue du chauffage central, etc.

Les nécessités de conservation de l'énergie ont imposé des modifications importantes dans la mise en projet des logements ayant pour but l'augmentation du degré de thermoisolation des bâtiments. La pratique de mise en projet de différents pays a montré que pour un dimensionnement rationel du point de vue économique des réseaux de distribution il est nécessaire à connaître exactement le niveau de la consommation actuelle, un calcul assez exact pour les suivants 5 ... 10 ans et un calcul moins précis à long terme.

Dans ce qui suit on présente l'étude de l'évolution de la consommation de l'énergie électrique d'un échantillon de 160 appartements de la ville de Jassy, dans la période 1985—1990, dans laquelle ont été appliqués plusieurs tarifs et différentes restrictions de la consommation de l'énergie dans le secteur domestique.

2. Le choix de l'échantillon d'étude

L'échantillon concerne un nombre de 160 appartements de la ville de Jassy. Les appartements sont différents suivant la zône (centrale, intermédiaire), le nombre des pièces, le nombre des personnes, le type du bâtiment. Tous les appartements sont munis avec chauffage central.

La consommation de l'énergie électrique a été surveillée pendant 16 périodes (Tableau 5).

Les tarifs appliquées dans la période 1985—1990 (Tableaux 2 ... 4) sont les conséquences des Décrets 240/1982, 27/1987, 272/1987, respectivement, 3/1990. Les premiers trois décrets sont très sévères et restrictifs. Le Décret 2/1990 a établi le tarif de 0,65 lei/KWh pour toute la quantité d'énergie électrique consommée.

Tableau 4

Les tarifs de 1987/II (concernent la consommation mensuelle)

Type de l'appartement	Le nombre de KWh ayant le tarif de 0,65 lei/KWh		Les tarifs des dépassements de la consommation
	novembre—février	mars—octobre	
studio	22	15	10%—2 lei/KWh
à 2 pièces	30	20	10%—3 lei/KWh
à 3 pièces	35	25	
à 4 pièces	42	30	

Tableau 5

Les périodes d'étude de l'évolution de la consommation de l'énergie

	Les périodes d étude	Les tarifs
1	18. 09. 1985—14. 12. 1985	tarif de 1982
2	14. 12. 1985—13. 03. 1986	
3	13. 03. 1986—19. 06. 1986	
4	19. 06. 1986—16. 09. 1986	
5	16. 09. 1986—10. 12. 1986	
6	10. 12. 1986— 8. 02. 1987	tarif de 1987/I
7	8. 02. 1987—26. 03. 1987	
8	26. 03. 1987—27. 06. 1987	
9	27. 06. 1987—19. 09. 1987	
10	19. 09. 1987—13. 11. 1987	
11	13. 11. 1987—21. 12. 1987	tarif de 1987/II
12	21. 12. 1987—15. 03. 1988	
13	15. 03. 1988—13. 10. 1988	
14	13. 10. 1988— 3. 03. 1989	
15	3. 03. 1989—19. 10. 1989	
16	19. 10. 1989—24. 01. 1990	tarif de 1990

3. Les résultats de l'expérimentation

Dans la Fig. 1 sont présentées les évolutions des consommations de l'énergie électrique pour différents types d'appartements. Normalement la consommation de l'énergie électrique est fonction aussi des saisons de l'année. Elle a la tendence de suivre périodiquement une augmentation assez forte pendant les mois froids d'hiver.

On constate, dans tous les cas, une augmentation de la consommation avec le nombre des personnes/logement.

Tableau 1

La structure de l'échantillon d'expérimentation

Catégorie de la zone de la ville	Le nombre des appartements de l'échantillon	Le nombre des pièces/app.	Le nombre des personnes/appartement						
			1	2	3	4	5	6	7
intermédiaire	25	studio	10	10	5	—	—	—	—
	50	2	—	17	19	9	5	—	—
	50	3	3	22	17	8	—	—	—
centrale	35	4	—	—	2	2	17	7	7

Tableau 2

Les tarifs de 1982 (concernent la consommation mensuelle)

Type de l'appartement	Le nombre de KWh ayant le tarif de :			
	0,65 lei/KWh	0,75 lei/KWh	0,85 lei/KWh	1 leu/KWh
studio	450	50	50	la différence
à 2 pièces	650	50	100	la différence
à 3 pièces	700	70	70	la différence
à 4 pièces	850	85	100	la différence
à 5 pièces	900	100	100	la différence

Tableau 3

Les tarifs de 1987/I (concernent la consommation mensuelle)

Type de l'appartement	Le nombre de KWh ayant le tarif de :			
	0,65 lei/KWh	1 leu/KWh	2 lei/KWh	2,5 lei/KWh
studio	50	2	2	la différence
à 2 pièces	70	4	4	la différence
à 3 pièces	78	4	4	la différence
à 4 pièces	95	5	5	la différence

De même on constate une forte diminution de la consommation de l'énergie électrique due à l'application des tarifs assez sévères (Décrets 27, 272/1987).

La conséquence de l'application du tarif de 0,65 lei/KWh pour toute la quantité d'énergie consommée (Décret 3/1990) constitue le début d'une période d'augmentation de la consommation de l'énergie électrique.

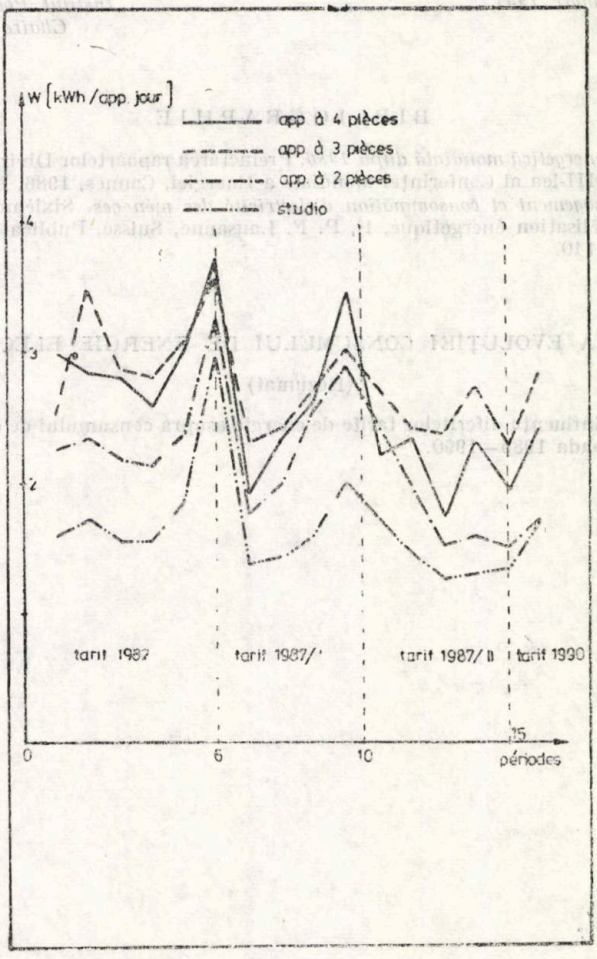

Fig. 1. — L'évolution de la consommation de l'énergie électrique.

4. Conclusions

Dans aucune pays du monde on ne saisit encore une diminution de la consommation de l'énergie électrique dans le secteur domestique. Pour les États Unis d'Amérique est prévue une saturation vers les années 2000.

La consommation de l'énergie électrique dans le secteur domestique en Roumanie se trouve encore à un niveau bas ($\approx$ 900 KWh/app. année). On peut prévoir une augmentation de cet indice dans l'avenir.

Le tarif fixé en 1990, de 0,65 lei/KWh, peut s'avérer inopportun pour une politique à longue terme dans le domaine énergétique.

Il est nécessaire de surveiller au niveau national l'évolution de la consommation de l'énergie dans tous les secteurs, dans le cadre d'une politique cohérente.

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Chaire d'Énergétique

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ASUPRA EVOLUȚIEI CONSUMULUI DE ENERGIE ELECTRICĂ

(Rezumat)

Se prezintă influența diferitelor tarife de energie asupra consumului de energie în sectorul domestic, în perioada 1985—1990.

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Tomul XXXVII (XLI), Fasc. 1—4, 1991
Secţia III
ELECTROTEHNICĂ. ENERGETICĂ. ELECTRONICĂ

RECENZII BOOK REVIEWS

REICHL H. und Mitautoren, *Halbleitersensoren, Prinzipien, Entwicklungsstand, Technologien, Anwendungsmöglichkeiten*. Expert Verlag, Ehningen bei Böblingen, 1989, 412 S., ISBN 3-8169-0221-9.

Sensoren stellen Schlüßelbauelemente für Problemstellungen aus dem Bereichen der Meß-, Steuer-, Regelungs- und Automatisierungstechnik dar. Für ihre Herstellung werden eine Vielzahl von Materialien und Technologien benötigt.

Halbleiter können aufgrund ihrer Bänderstruktur zur Umwandlung von einer überaus großen Vielzahl von physikalischen und chemischen Größen in elektrische Signale eingesetzt werden. Als Materialien werden bevorzugt Si, Ge, GaAs, InSb. InAs, SnO_2 und ZnO eingesetzt.

Das Buch ist in der Reihe „Kontakt & Studium" publiziert, die nicht nur lehrgangsbegleitende Funktion hat, sondern auch alle Voraussetzungen für ein effektives Selbststudium erfüllt. Die Breite des dargestellten Gebiets kann man gut den Kapitelüberschrift entnehmen: 1. Die Grundprinzipien moderner Halbeitersensoren (H. Reichl); 2. Herstellungstechnologien für Halbleitersensoren (E. Obermeier); 3. Halbleiter — Temperatursensoren (G. Kowalski); 4. Piezoresistive Silizium — Drucksensoren (J. Binder); 5. Mikromechanishe Silizium — Beschleunigungssensoren (V. F. Vilimek); 6. Magnetfeld — Positionssensoren (W. Flossmann); 7. Fotosensoren mit Glasfaseroptiken — Vielfältige Einsatzmöglichkeiten in der Industrie (W. Schultz); 8. Halbleitersensoren zur Bildwandlung (R. J. Ahlers); 9. Gassensoren (H.-P. Hübner); 10. Meßschaltungen und Meßsignalverarbeitung für Sensoren (H.-R. Tränkler).

Der Inhalt des Buches und die Art und Weise der Behandlung spiegeln die Hochschul- und Forschungserfahrung der Autoren wieder, die um Anschaulichkeit bemuht sind und durch Querverbindungen, zahlreiche und sorgfältig ausgewählte Einzelheiten und treffende Illustrationen fördern.

Das Buch ist geeignet für Studenten automatisierungs- und meßtechnischer Fachrichtungen, in der Praxis tätige Entwicklungsingenieure für Automatisierungstechnik, Projektanten und Technologen.

M. Voicu

KOWALSKI H.-J., *Berechnung und Aufbau aktiver RC-Filter*. Militärverlag der DDR, Berlin, 1988, 128 S., ISBN 3-327-00552-4.

Aktive *RC*-Filter haben in den letzten Jahren durch die Fortschritte der Mikroelektronik außerordentlich an Bedeutung gewonnen. International wird ein breites Spektrum derartiger Filtermodule angeboten. Die Hauptanwendungen aktiver *RC*-Filter liegen im Frequenzbereich von Millihertz bis zu einigen hundert Kilohertz.

Anliegen des Autors ist es, dem Leser anwendungsbereites Material zum Entwurf aktiver *RC*-Filter und praktisch erprobte Schaltungen vorzulegen. Die Breite des dargestellten Gebiets kann man gut den Kapitelüberschrift entnehmen:

1. Einführung ; 2. Filter funktionen ; 3. Anforderungen an die passiven und aktiven Bauelemente für die Realisierung aktiver *RC*-Filter ; 4. Der praktische Aufbau aktiver *RC*-Filter ; 5. Elektronisch steuerbare *RC*-Filter ; 6. Aktive *RC*-Filter für den HiFi-Amateur.

Besonders wendet das Buch an Studenten im zweiten Studienabschnitt und Diplomanden sowie an Wissenschaftler und Ingenieure in Forschung, Entwicklung und Produktion, die für die praktische Anwendung der aktiven *RC*-Filter sind.

M. Voicu

BOLTJANSKI G. W., *Optimale Steuerung diskreter Systeme*. Geest & Portig K.-G., Leipzig, 1976, 326 S.

Dieses Buch stellt einen ersten Versuch der systematischen Darstellung der Theorie der diskreten optimalen Steuerung dar. Es behandelt nur rein mathematische Fragestellungen. Aber auch bezüglich der mathematischen Theorie diskreter optimaler Prozeße erhebt das Buch keinen Anspruch auf Vollständigkeit. Zum Beispiel werden diskrete Prozeße mit einer unendlichen Anzahl von Schritten, Fragen der Approximation stetiger optimaler Prozeße durch diskrete Prozeße und numerische Methoden für Aufgaben der diskreten Steuerung überhaupt nicht betrachtet.

Zwei Methoden liegen der Konzeption des vorliegenden Buches zu Grunde. Die erste ist mit Fragen der Trennbarkeit eines Systems konvexer Kegel verbunden. Diese Methode wird in wesentlich allgemeinerer Form dargestellt. Die zweite Methode ist mit Begriffen aus der Topologie verknüpft (und zwar mit Schnittheorie).

Das Buch enthält vier Kapitel : 1. Aufgabenstellung und Charakter der Resultate ; 2. Elemente der Theorie konvexer Mengen ; 3. Extrema von Funktionen ; 4. Optimalitätskriterien für diskrete Prozeße. Die Besonderheit des Buches liegt darin, daß es aus vier im wesentlichen einzelnen, kleinen Büchern besteht, die zusammen ein einheitliches Ganzes bilden.

ANISHCHENKO V. S., *Dynamical Chaos in Physical Systems — Experimental Investigation of Self-Oscillating Circuits*. Teubner, Leipzig, 1989, 212 pp., ISBN 3-322-00713-8.

This book is published in the series „Teubner — Texte zur Physik" (vol. 22) and discusses the results of numerical and physical experiments to investigate the origin, structure and properties of strange attractors in electronic systems of small dimension. If in the neighbourhood of homoclinic structure stable regular trajectories are absent (points of equilibrium, limit cycles, tori), then the chaotic regime of oscillations is realized, represented mathematically by a strange attractor.

In this book the chaotic oscillations in simple radiophysical systems are investigated, which are represented by quasi-hyperbolic attractors. The principal theoretical concepts and results were reported in the book *Dynamical Chaos — Basic Concepts* (Teubner, Leipzig, 1987), by the same author, while in the present book the main attention is devoted to the experiments. The topic covered follow a logical sequence through four chapters whose titles are : 1. Quasi-attractors in the expanded neighbourhood of special trajectory of separatrix loop type; transition to chaos and its properties in generator with inertial nonlinearity ; 2. Mechanisms of formation ; structure and properties of torus-attractor ; 3 Transition to chaotic oscillations in distributed systems ; possibility of dynamic description ; 4. Influence of fluctuations on properties of quasihyperbolic chaos.

A special feature of the treatment consists in the detailed comparison of numerical experiments results with the data of measurements made on real devices adequate to corresponding mathematical models.

The book is highly recommended to electronics engineers and any other profesionals (physicists, systems engineers, applied mathematicians) interested in the investigation of dynamical chaos in physical systems.

M. Voicu

STECHER R., *Messung von Zeit und Frequenz*. VEB Verlag Technik, Berlin, 1990, 198 S., ISBN 3-341-00661-3.

The book of dr. sc. techn. Rudolf Stecher, *Messung von Zeit und Frequenz*, is part of the famous collection „Messtechnik", which polarized round it, a big number of great specialists, forming a High School of Metrology, with high scientific and didactical level.

Written in a very concise and clear style, this work offers for the reader a great volume of up-to-date practical and theoretical information, dealing with the field of time, frequency and phase measurement.

So, in the first and the second chapter, the author debates the main concepts about the definition of the second and the timescales (astronomic, atomic, international, international coordinated time) ; in the chapters 3 and 4 he presents the atomic time-standards (with cesium, hydrogen and rubidium), with a deep analysis of the performances and the limits of them.

In the chapters 5, 6 and 7 — the central part of the book — the author deals with the main methods and devices for measuring time, frequence and phase, with emphasis on the digital frequency and chronometers. Very interesting is a method of measuring optical frequencies.

At last, in chapters 9 and 10, are presented the principal methods of transmiting the frequency and time units : by cable, by radio and by satelitte, and the principal stations and chanels (via satelitte) for the distribution of the precise hour.

Through its contents and richness of topical informations, the present book is a very useful working aid for the specialists from industry and high education.

M. Antoniu

BUTZ U., HEINZE D., NEUMANN I., *Softwareentwicklung für die Projektierung von Elektroenergeanlagen*. 1. Auflage, Verlag Technik GmbH, Berlin, 316 S.

Die Autoren haben ein Buch erarbeitet, das die bereits zur Softwareentwicklung erschienenen Titel in der Weise ergänzt, daß hier eine Verbindung zu einer Fachdisziplin hergestellt wird, nämlich zur Projektierung von Elektroenergieanlagen. Die Projektierung ist ein Auswahl- und Dimensionierungsprozeß, der sich wiederholende Abläufe aufweist und daher für eine rechentechnische Aufbereitung geeignet ist.

Nun geht es darum, eine Verständigungsbasis zu schaffen, auf der die anlagentechnische zur rechentechnischen Seite in Beziehung gesetzt werden kann und umgekehrt. Diese Verständigungsbasis ist das sogenannte konzeptuelle Modell.

Die Modellierung des Projektierungsprozeßes wird in allen Schritten gezeigt. Die Autoren stützen sich dabei auf das sogenannte Drei-Ebenen-Modell (Informationsbasis, Verarbeitungsebene, Prozeßsteuerebene).

Dazu werden erforderliche Beschreibungsmittel, wie Graphen, Petri-Netze, semantische Netze, endliche Automaten, in ihrer Anwendung auf diesen Projektierungsprozeß vorgestellt.

Interessenten : Entwicklungs- und Projektierungsingenieure ; Studenten.

BIMBERG D. (Federführer), AMENDE W., KREUTZ E. W., SEILER P., TREUSCH M.-G., *Materialbearbeitung mit Lasern. Grundlagen und Anwendungen.* Reihe Kontakt u. Studium, Bd. 343, Expert-Verlag GmbH, Ehningen bei Böblingen, ISBN 3-8169-0335-5.

Das Buch vermittelt einen flächendeckenden Überblick über die Praxis der Materialbearbeitung mit Lasern. In einzelnen Kapiteln werden detailliert Schweißen, Schneiden, Bohren, Beschriften, Fugen, Löten, Härten, Umschmelzen, Legieren und viele andere Verfahren von Spezialisten der jeweiligen Anwendungsgebiete erörtert. Viele Bildbeispiele veranschaulichen die unterschiedlichen Einsatzmöglichkeiten, die von den Bereichen der Feinwerktechnik und Elektronik bis in den Flugzeugbau reichen. Bei der Diskussion der großen Zahl von Einzelbeispielen wird auf Wirtschaftlichkeitsaspekte besonderer Wert gelegt. Der mit dem Laser als Lichtquelle noch nicht vertraute wird in einem einleitenden Kapitel über die Grundprinzipien aufgeklärt. Es folgt ein Überblick über die äußerst unterschiedlichen Typen von Lasern, die in der Fein- und Grobbearbeitung Verwendung finden. Schließlich werden die Grundlagen der Wechselwirkung von Laserstrahlung mit unterschiedlichen Materialien so eingehend behandelt, daß dieses Buch selbst an Hochschulen und Fachhochschulen als Grundlage für Lehrveranstaltungen dienen kann.

ELLISON G. N., *Thermal Computations for Electronic Equipment.* Robert E. Krieger Publishing Company, Malabar, Florida, USA, 1989, 397 pp., ISBN 0-89464-401-7.

This book is a reprint edition of the original edition from 1984. It was very much appreciated by the readers.

The book contains only those aspects of heat transfer in electronic equipment with which the author considered himself to be fully acquainted and that he have applied to practical design problems. He have confined the material totally to conduction, radiation, and convection by natural or forced airflow. These are the heat transfer mechanisms of great interest to designers of electronic test instruments, computer systems, and other ground-based equipment.

The digital computer programs included here (TAMS and TNEFTA written in FORTRAN IV) are offered as serious analytical tool.

The book is a reference scientific and engineering book for the users of diverse academic preparation.

N. Damean

HOCK A., PAULI P u.a., *Antennentechnik.* Expert Verlag, Deutschland, 1989, 268 S., ISBN 3-8169-0400-9.

The present work is one item of a large collection „Mess- und Prüftechnik" (Measuring and Testing Techniques) published by the Expert Verlag book company. The collection is intended to all the specialists who really want to be kept well-informed in their field of activity. There are underlined, in the preface, the aims of the „continuous education" and its importance in the context of everyday changes. This book is even more useful, as most of the electrical engineers having to maintain and use antenna installations received just elementary theoretical background during their academic studies.

Keeping in mind the above statements the authors present in the first chapter of the book the antenna theoretical fundamentals and the antenna main parameters : reciprocity theorem, polarisation, radiation pattern, radiation resistance, effective height, etc. The half wave dipole is completely presented, as well as linear arrays with half wave dipoles.

The second chapter deals with the general methods and the specific problems encountered in the antenna measurement technique. There are presented the

external influences, the errors and the final results delivering forms for the antenna input impedance measuring. For the radiation pattern measuring, the plane wave criterion is firstly stated and, then, the reflected wave elimination methods, the measurement precision and some near field measurement methods are analysed. The antenna gain measurement is, also, presented.

The third chapter is devoted to a special class of antennas — the radar ones. It comprises a general presentation of the microwave antennas (especially, horn antennas) and waveguides, as well as, specific problems regarding the primary and secondary radiator, pencil beam antennas, air-borne antennas, phased array antennas, etc.

The following chapter presents the problems of the point-to-point communication antennas : gain, cross area, frequency band, front-to-back ratio, mechanical tolerances, de-focusing, etc.

The fifth chapter deals with the radio antennas for the long, medium and short wave ranges. It is divided in two main parts : surface wave antennas and ground wave antennas. The transmitting antennas are mainly discussed, but there are some considerations upon the reception antennas, too.

The sixth chapter analyses the VHF, UHF and SHF antennas comprising the electrically short dipole, the half wave dipole, the full wave dipole, the Yagi array antennas, the log periodic antennas, the spiral-conical antenna, the horn antennas, the reflector antennas, etc.

The seventh chapter deals with the TV transmitting antennas (including associated sound transmitting). Firstly, the noise and perturbations problems are analysed ; then, the most used antenna types are described : single channel antenna, multiband antenna, Yagi antenna. At the end of the chapter the mechanical construction and installation of an antenna and some other auxilliary means are presented.

The eight chapter describes the problems encountered in the industrial process of an antenna manufacturing. As these problems are specific for a given antenna type they are grouped in separate parts for the microwave antennas, the radio transmitting antennas (LW, MW, SW and USW), the auto antennas, the active antennas.

The last chapter is entirely devoted to the active reception antennas. The definition of an active antenna is given, its necessity is underlined and its specific problems are analysed, such as signal to noise ratio, frequency band, non-linear effects, etc.

Excellent references are given at the end of the book. They are in a great number (over 130), up-to-dated and grouped as the book's chapters are.

Every chapter is written by a specialist in the field, so that the book practically has 8 authors. Thus, a professional presentation of every chapter is made and a good mixture of theory and praxis results.

By the subject actuality, by the manner of the presentation and by the valuable knowledge it contains the book become a very necessary and useful tool for any electrical engineer or student in the field.

I. Bogdan

STEINKE E., SCHOLZ K.-P., *Ton und Bild, Aufnahme- und Wiedergabetechnik Lexikon*. Verlag Technik, Berlin, 1990, 246 S., ISBN 3-341-00892-6.

Um dem allgemeinen Informationsbedürfnis entgegenzukommen, wurde dieses kleine Lexikon geschafen. Es erklärt häufig verwendete Begriffe des Fachgebiets in knapper Form. Durch zahlreiche Strichzeichnungen wird der Text ergänzt. Neben spezifische Termini, die erst mit der Ton- und Bildtechnik entstanden sind, wurden auch Begriffe aufgenommen, die aus der Elektrotechnik, Mechanik, Elektronik und Optik stammen, aber mittlerweile in verstärktem Masse in die Ton- und Bildtechnik Eingang gefunden haben und für das Verständnis gleichermassen wichtig sind.

Der Band ist vor allem für jene gedacht, die sich erstmalig mit Ton- und Bildtechnik befassen, doch werden auch Techniker, Leitungskader und Facharbeiter, aber auch Ökonomen, die mit Vorbereitung oder dem praktischen Einsatz von dieser Technik in Berührung kommen, Nutzen aus diesem Band ziehen können.

HARRIS E. K., ALBERT A., *Survivorship Analysis for Clinical Studies*. Marcel Dekker Inc., New York, 1991, XVI+200 pp. and a diskette ; ISBN 0-8247-8400-6.

The book is vol. *114* of series for statistics : textbooks and monographs, edited by Department of Statistics, Southern Methodist University Dallas, Texas.

This volume will be of interest to all who use statistics in their work and those who are concerned with statistical estimation theory and it applications : biostatisticians, biomedical engineers, clinicians, pharmacologists.

The book contains seven chapters. The first three chapters are devoted to the estimation and comparision of survival curves with calculation of the standard errors of estimates of survival rates. Chapter 3 analysis the comparison of two or more etimated survival curves. Chapters 4 and 5 describes methods for quantifying the effects of covariantes on survival rates, graphical tests for proportional hazards and nonproportional hazards. Chapter 6 examines recent contributions to analysis of the effects of time-dependent covariant. The last chapter presents various computer programs currently available to apply the methods presented.

Attached to this book is a diskette. The diskette contain two programs for compuntig confidence bands for survival curves estimated by the K a p l a n - M e i e r method.

C. Voiniciuc

GLASS L., MACKEY M. C., *From Clocks to Chaos. The Rhythms of Life*. Princeton University Press, 1988, XVII+248 pp.

The book is divided in nine chapters and a mathematical appendix. A brief outline is given below.

Chapter 1. *Introduction : The Rhythms of Life*. This chapter presents an introduction to most of what is to follow, discussing the mathematical models for biological oscillators, the perturbation of physiological rhythms, and introduces the concept of dynamical disease.

Chapter 2. *Steady States, Oscillations and Chaos in Physiological Systems*. The concepts of steady states, oscillations, chaos, limit cycles, phase planes, bifurcations and stability as applied to biological systems are introduced, as well as another class of widely used models : the finite difference equations.

Chapter 3. *Noise and Chaos*. The authors tackle the task of further describing the mechanism of chaotic phenomena and introduce other useful notions, such as strange attractors, dimension, L y a p u n o v numbers and present a comparative discussion of noise versus chaos.

Chapter 4. *Mathematical Models for Biological Oscillators*. This chapter presents the main categories of models that have been proposed for biological oscillators : the generation of pacemaker oscillations, mutual inhibition and sequential disinhibition as central pattern generators and the negative feedback as a control mechanism. The authors show that such systems can lose stability and manifest oscillations.

Chapter 5. *Initiation and Termination of Biological Rhythms*. The authors discuss in this chapter the experimentally observed transitions between oscillatory and nonoscillatory states and present some explanatory hypotheses. The cases of soft and hard excitation as well as the possibility of annihilating the oscillation with a single stimulus are also studied.

Chapter 6. *Single Pulse Perturbation of Biological Oscillators*. This chapter deals with the experimental and theoretical results of single stimuli applied to biological oscillators and the theory of phase resetting in various fields.

Chapter 7. *Periodic Stimulation of Biological Oscillators*. The authors present the main experimental observations concerning the phase locking phenomena and then develop a theoretical analysis of such phenomena in integrate and fire and limit cycle models.

Chapter 8. *Spatial Oscillations*. This chapter discusses the wave propagation in one, two and three dimensions as well as experimental and theoretical aspects of fibrilations.

Chapter 9. *Dynamical Diseases.* The concept of dynamical disease, their mathematical models and the way these models could be used for diagnosis and therapy design are presented in this last chapter.

Mathematical Appendix. The authors present the basic mathematical techniques required to understand the book in depth and to be able to construct some models for particular biological systems. Finite difference and differential equations are discussed separately.

The book is written in a clear manner, which must have been a real challenge, given the fact that the treatment of chaotic dynamics usually requires a strong mathematical background while there would be readers of this text generally lack exactly such a background. Therefore, having succeeded an accurate presentation of chaos in biological systems without using too complex a mathematical framework must be considered solely as the merit of the authors. Perhaps the greatest achievement of this book is its demonstration that physiology can and should benefit from the use of nonlinear mathematical modelling in order to better understand dynamic phenomena in biological systems.

The book has a good graphical presentation, with many illustrations and graphs. The authors also provide a rather rich bibliography.

Biologists, physicians, students and researchers in these fields should find this book thought-provoking and worth reading. Mathematicians and researchers in chaotic dynamics might find interesting the numerous biological examples presented. Although a comprehensive mathematical background is not required in order to generally understand the book, it is a prerequisite for an in-depth assimilation of chaotic dynamics.

I. Cr. Bujor

HOLDEN A. V. (Ed.), *Chaos.* Princeton University Press, 1986, VIII+324 pp., ISBN 0-691-08424-6.

The book is a collection of 15 papers grouped in six parts. Their brief outline is given below.

Part I. *Prologue.* This chapter contains two introductory papers whose aim is to familiarize the reader with the world of chaotic dynamics discussing, first, the utility of these concepts and then, various types of attractors.

Part II. *Iterations.* This chapter contains two papers by the same author discussing various aspects and properties of iterative maps in one and two dimensions.

Part III. *Endogenous Chaos.* This part contains five papers dealing with endogenous chaos in such various fields as lasers and optical resonators, ecology and epidemiology, cellular metabolism and physiological systems. It also contains a paper dedicated entirely to the famous Lorenz equations.

Part IV. *Forced Chaos.* This part contains three papers dealing with forced nonlinear oscillators, chaotic cardiac rhythms, and chaotic oscillations and bifurcations in squid giant axons.

Part V. *Measuring Chaos.* This part contains two very interesting papers aimed, respectively, at determining quantitative aspects of chaotic systems using Lyapunov exponents and at estimating the fractal dimensions and entropies of strange attractors.

Part IV. *Epilogue.* This part contains a rather amusing but nevertheless interesting paper by D. E. Rossler which shows the vigour of this field of nonlinear dynamics.

This book is a collection of papers dealing with various aspects of chaotic dynamics, from central concepts like iterative maps, Lyapunov exponents, strange attractors and their fractal dimensions, to their application in such fields as ecology, epidemiology, cellular metabolism, cardiology, electronic oscillators, lasers and optical resonators.

It has a good graphic presentation with many figures and graphs. Each paper has a list of references.

This book addresses a potentially large audience, ranging from biologists and physicians to physicists and mathematicians. Although the book is not written at an introductory level, its contents are accessible to readers less familiar with nonlinear dynamics but with an adequate mathematical background.

I. Cr. Bujor

SCHOLZ K. P., STEINKE E., *Lexikon Rundfunk- und Fernsehempfangstechnik*. VEB Verlag Technik, Berlin, 1990, 256 S., ISBN 3-341-00543-9.

This dictionary of Radio and TV Communication techniques contains about 1000 useful words. Some of the explanations are quite concise but many notions are explained in a thorough manner. The explanations are very clear and up to the point. Some notions are illustrated with schematics and drakings. Some abbreviations are also introduced and explained.

In conclusion, an useful book, not to be left out of our library.

N. D. Alexandru

WINFREE A. T., *When Time Breaks Douwn. The Three-Dimensional Dynamics of Electrochemical Waves and Cardiac Arrhythmias*. Princeton University Press, New Jersey, 1987, XIV+340 pp., ISBN 0-691-08443.

Arthur T. Winfree is Professor in the Department of Ecology and Evolutionary Biology and the Program in Applied Mathematics at the Universitv of Arizona.

Professor Winfree proposes us an original mode for investigate the breakdown of rhythmic biological timekeeping. He applies scientific knowledge from different disciplines as : biology, ecology, psychology, traditional medicine, anatomy, computer science, chemistry.

The book contains four parts with ten chapters. A multidimensional appendix, a glossary, references, an index of citations and a subject index complete the book.

The first part describes biological rhythms of the mammalian heartbeat. their electrical signature and the susceptibility of the rhythm to lethal disruption by an unfortunately timed stimulus. The second part presents the electrical conversion to lethal arhythmia by a singular stimulus. Part III studies the spatially and temporally rhythmic and periodic excitation of heart. The author illustrates his fundamental ideas by coloured diagrams and the phase of a rhytm is indicated by a corresponding color with computer graphics. Professor Winfree finishs his book with question mark „What's next ?" and discoveries to come at the last part.

Winfree's book constitues an excellent documentary and inspiration source for the general scientist.

C. Voiniciuc

BAYLIN F., GALE B., *Ku-Band Satelite TV Theory, Installation and Repair*. Baylin/Gale Productions, 360 pp., ISBN 0-917893-06-9.

The book provides a comprehensive exploration of the theory and practice underlying Ku-band satellite television. The interested reader can be anyone, even without a technical background, as the material is clear understandable.

This book contains 9 chapters and 6 appendices. The first chapter. „The Basics of Satellite Communications", introduces the fundamentals. The second chapter, „Component Theory and Operation" treats antennas, antenna mounts, actuators,

feedhoms, low-noise block downconverters, downconversion methods, coaxial cable and connectors, satellite receivers, modulators, TV receivers and monitors and finally monophonic and stereo audio reception.

The third chapter, „Design Characteristics of Ku-band Systems", treats the influence of various factors on the design, such as path loss, antenna gain, system noise temperature, etc.

Very interesting is chapter IV, „Selecting Satellite TV Equipment", which should be most attractive to anyone. Also, chapter V, „Installing Ku-band Systems", which is very well written.

Chapter VI, „Retrofitting C-band with Ku-band Systems", deals with performance requirements for Ku-band components and installation procedures.

Chapter VII, „Multiple-Receiver Satellite TV and Distribution Systems", is a survey of the basic components followed by design consideration of the headend and distribution systems. In chapter VIII, „Worldwide Ku-band Satellite Television", some data about various regions and maps are provided.

Finally, chapter IX, „Troubleshooting and Repair", concludes the book. In the appendices, basic data about satellite communication systems are provided.

In conclusion, a very interesting and useful book which should appeal to anyone interested in satellite TV reception.

N. D. Alexandru

SCHÖNFELD R., *Digitale Regelung elektrischer Antriebe.* VEB Verlag Technik, Berlin, 1987, 240 S., ISBN 3-342-00303-7.

The stage of development of the integrated microelectronic circuits and the advances in software have allowed the improvement of the numerical control and setting of the electrical drives of the system.

These problems — structured into 10 chapters — are dealt with in this look.

The first chapter discusses the numerical processing of the signals in electric drives by referring to the quantification of the continuous values, coding, filtration of the numerical signals, numerical — analogue and analogue — numerical conversions, transmission of the numerical signals.

The second chapter presents the programming of the control and setting devices by referring to microcomputer structures. The devising and the optimization of the numerical control more loops — setting structures are taken into account.

The fourth chapter presents the numerical control and setting of the speed of the electrical driving systems, of the position, the control of the movement on more axes, the control of the step-by-step drive.

The control and the setting of the current in the case of the continuous current driving are discussed in the fitfh chapter.

The sixth chapter deals with the control and setting of the current at the forced commutation switch.

The control according to the field of the three-phase machine is presented in the seventh chapter.

The eighth chapter discusses the numerical setting of speed and position.

The setting and the optimal control of the movement by taking into account the rigid and flexible drives are described in the ninth chapter, while the tenth chapter presents hardware and software problems, the service and the maintenance of the numerical setting systems.

The bibliography contains 250 references and is followed by an alphabetical index.

This book is intended for the specialists working in the domain of designing, controlling and servicing the adjustable electrical driving systems.

Simona Caba

HOMMEL G., *Prozessrechensysteme '91*. Springer Verlag, Berlin-Heidelberg-New York, 1991, 449 S., ISBN 3-540-53808-9 ; 0-387-53808-9.

The volume contains the works of the session „Prozessrechensysteme '91 Automatisierungs und Leitsysteme in den neunziger Jahren" which was held in Berlin on February 1991. This session has belonged to the tradition of the scientific sessions having as subject the computers of process set up in West Germany since 1974. Since 1978, in Stuttgart here has been adopted as title of the session „Prozessrechensysteme" in order to emphasize that in computer leading systems the leading central computer is substituted for complex systems of process computers. Thus, distributed configurations of calculation are obtained which working in real time and with time division having as advantages the safety in working and tolerance against errors.

The works presented in the volume deal with the definition of the requirements imposed to the computer leading system in projecting and configuration determination fields, system architecture and communication between levels, software technologies and applications. The present technical state, the perspectives concerning the conversational technique, the hardware systems principles in real time, expert systems and systeme tolerant to error are all reviewed.

The volume contains 39 works classified in 13 sections. This volume belongs to „Informatik-Fachberichte" collection.

Simona Caba

BÄSSLER R., DEUTSCH A., *Nachrichtennetze*, VEB Verlag Technick, Berlin, 1989, ISBN 3-341-00575-3.

Die sehr schnelle Entwicklung des Gebietes Fernmeldewesen ist ein Merkmal unserer Zeit. Um kalifizierte Kräfte schulen zu können ist die entsprechende Fachliteratur besonders wichtig. Das Werk nimmt sich vor eine zusammenfassende Quelle betreffs der Nachrichtennetze zu bieten. Eine wichtige Aufgabe falls man die entsprechende mangelnde Literatur betrachtet.

In einer gut gewählten Reihenfolge werden, im Grunde, folgende Kapiteln präsentiert : Grundsätze der Nachrichtennetze ; Grundlagen zur Gestaltung und Plannung von Nachrichtennetzen ; Gestaltung und Plannung von Anschlusskabel- und Anschlussleitungsnetzen ; Gestaltung und Plannung von Orts- und Knotennetzen ; Gestaltung und Plannung nationaler Weitverkehrsnetze ; Verkehrstheorie.

Infolge des reichen Informationsinhalt und der Klarheit der Darstellung bildet das Buch ein wertvolles Instrument sowohl fürdijenigen die sicht mit der Projektierung von Fernmeldenetzen befassen, als auch für die Studenten der Hoch- und Fachschulen mit einschlägigen Fachrichtungen.

M. Antonescu

EIGLER H., *Mikroelektronische Filter*. Verlag Technik GmbH, Berlin, 1990, 332 S., ISBN 3-341-00312-6.

Mikroelektronische Filter sind Festkörperschaltkreise oder Funktionalelemente. Sie verwirklichen alle notwendigen Operationen für die Filterung und Verarbeitung elektrischer und auch optischer Signale in einem Volumen von einigen zehn bis hundert Kubikmillimetern. Heute bestimmen sie, wie modern und wirtschaftlich Systeme der Nachrichtentechnik, Messtechnik, Steuer- und Regelungstechnik sind. Mikroelektronische Filter überdecken das gesamte von der Informationstechnik genutzte Frequenzband ; bei Einbeziehung der Lichtwellenleiter verläuft das Frequenzband bis in den Mikrometerbereich. Diese Filter sind auch ein markantes Beispiel für die durch sie möglichen umsetzungen theoretischer Arbeiten und Entwicklungen bei der zeitdiskreten Signalverarbeitung.

Die mit der theoretischen Vorbereitung, Entwicklung und Einführung zu lösenden Filteraufgaben nehmen bis heute und sicher in Zukunft einen breiten Raum in der Spezialliteratur ein, jedoch fehlt eine gemeinsame, möglichst einheitliche Darstellung der theoretischen Grundlagen, Wirkprinzipien, Anwendungsbereiche. Diese Lücke schliesst das Buch. In ihm werden : a) Schalterkondensatorfilter ; b) ladungsgekoppelte Filter ; c) Digitalfilter und Signalprozessoren ; d) Streifenleitungsfilter; e) akustische Oberflächenwellenfilter; f) optische Filter behandelt. Hierzu werden ein Einblick in die Mikroschaltungstechnik gegeben und die allen Filtern gemeinsamen systematischen Grundlagen aufgezeigt.

Besprochen werden weiterhin das physikalische Prinzip, die jeweiliger Anwendungen, Vor- und Nachteile, mögliche Perspektiven und die Grundlage für den Filterentwururf. Exemplarische Beispiele vertiefen den Praxisbezug.

Das Buch wendet sich nicht nur am den ausgesprochenen Filterfachmann. Es wurde für Ingenieure, Wissenschaftler, Studenten und für all diejenigen geschrieben, die sich für die neue Möglichkeit der Signalverarbeitung interessieren.

BÖDEKER K., GRIGOLEIT L., *Fachausdrücke und ihre Definitionen.* Verlag Technik, Berlin, 1990, 88 S., ISBN 3-341-00853-5.

In mehr als zweihundert Rechtsvorschriften sind etwa achttausend Fachausdrücke definiert, die heute von den Elektropraktikern verwendet werden.

Allerdings, kaum ein Fachkollege verfügt über alle diese Vorschriften und Gesetzblätter ; keiner kann gewiss sein, die Definitionen aller für ihn wichtigen Fachausdrücke zu kennen. Dies behindert die Ordnungsgemässe fachliche Arbeit.

In diesem Buch wird versucht, die für der Elektropraktiker des Elektroanlagenbaus wesentlichen Fachausdrücke zusammenfassen. Ihre Gliederung erfolgt nach Sachgebieten und nicht vorrangig nach dem Alphabet, um so auf die fachlichen Zusammenhänge unterschiedlicher Begriffe hinzuweisen.

Der begrenzte Umfang dieser Broschüre machte eine Beschränkung der Begriffe notwendig. Es wurden auch Ergänzungen erforderlich, weil nicht alle in der täglichen Arbeit vom Praktiker benutzten Fachausdrücke mit ihren Definitionen in den Rechtsvorschriften enthalten sind. Alle interessierten Fachkollegen werden gebeten, die Autoren darauf hinzuweisen, welches Sachgebiet und welcher Fachbegriff bzw. welche Definition künftig verändert, ergänzt oder gestrichen werden sollte. Beim Zusammenstellen der Begriffe zeigte sich, dass viele Fachausdrücke mehrfach und leider auch nicht immer gleichlautend definiert worden sind.

Inhaltich voneinander abweichende Definitionen wurden aufgenommen, sofern sie sowohl in der einen als auch in der anderen Form gebräuchlich sind.

Indem derartige Differenzen direkt gegenübergestellt werden, soll das Buch auch zu einer künftig besser koordinierten Arbeit auf diesem Gebiet beitragen.

Wir hoffen, dem Elektropraktiker hiermit ein Arbeitmittel zur Verfügung zu stellen, das ihm eine schnelle und bessere Information ermöglicht und bei Diskussionen zur Verfügung steht. Es soll zu einer einheitlichen Gestaltung und Anwendung der Fachausdrücke und so zu einer höheren Qualität der Arbeit verhelfen.

ROGERS K. S., KABRISKY M., *An Introduction to Biological and Artificial Neural Networks for Pattern Recognition.* A Publication of SPIE, The International Society for Optical Engineering, Washington, 1991, 220 pp.

The aim of S. K. Rogers and M. Kabrisby's book is to offer a brief introduction in a very recent and exciting research field : neural networks. The authors ave acquired an enormous experience into the treating of the pattern recognition problem using neural methods. Therefore they propose to emphasize some efficient tools for those researchers interested in pattern analysis for artificial perception tasks.

In addition to the eight chapters of the book there are also presented in the appendix some software examples, written in C, thought as being the best way to understand how neural nets appear to learn.

Firstly, a short introduction allows the reader to place the artificial neural networks domain. Chapter 2 deals with an historical overview by a review of the massive literature which place in perspective some of the currently popular models.

While chapter 3 contains an overview of a few biological exemples from general informations about neurons to an investigation into cortex' function, chapter 4 offers an explanation of how neural information processing systems work. Furthermore it is stated a definition of intelligence as applied to the interpretation of information.

The fifth chapter of the book discusses some of the principal characteristics of the most used artificial neural networks : reccurent nets feedforward nets and several possible applications such as target detection, expert systems, etc.

In the next chapter there are provided some insights into the pattern recognition problem and into the best known neural network learning algorithms. Some interesting details on simple and multilayer perceptrons, especially the backpropagation, are pointed out. The Kohonen self-organization feature maps serve as an example of an unsupervised learning paradigm. In the following there are discussed some classical applications. Another unsupervised learning algorithm, Grossberg's ART, is further developed in order to attack a problem of tactical target recognition. In the end of the chapter the authors examine the performances of Hopfield networks.

Chapter 7 proposes a way to apply the neural networks approach to some real-world problems : target recognition and speech recognition. The last chapter refers to a group of neural net development tools including backprop, hybrid networks, Hopfield associative memory, Gabor processing, Kohonen mappings, etc.

In appendix A it is explained how a MLP ANN using backward error propagation derived from common principles of gradient descent works. In appendix B there are presented the Kolmogorov results in trying to solve the thirteen problem of Hilbert. These results are used to justify why no more than three layers of perceptrons are needed to solve any discrimination task.

Valentin Drăgoi

ANDERSON Z. D. (Ed.), *Neural Information Processing Systems.* A Publication of AIP, American Institute of Physics, 1988, 871 pp., ISBN 0-88318-569-5.

This volume contains the proceedings of the first IEEE Conference on Neural Information Processing Systems held in Denver, Colorado, on November 8-12, 1987. The reader will be confronted with over 80 papers, each one of a very good quality.

The refered book offers a wide spectrum of neural applications into the information processing domain. Thus, one could be listed phoneme classification, pattern association, robotics, speech coding and perception, associative databases, connexionism as well as new hardware for neural nets implementation.

Everyone who read the contents will be able to find some articles which plan to place the artificial neural networks among the very used techniques of information processing. There are presented some extensions of the currently popular models in the context of the conference topics.

There have not been missed the biological approaches of neural networks by an investigation into cortex functions, hippocalampal neurons, vestibulo-ocular system, etc. Some of the papers offer an explanation of how neural information processing systems work. Furthermore it is stated a definition of intelligence as applied to the interpretation of information.

Some of the principal characteristics of the reccurent neural networks and their possible real applications such as learning state space trajectories, target detection, expert systems are also discussed.

There are also provided some insights into the pattern recognition problem and into the best known neural network learning algorithms. Some interesting details on simple and multilayer perceptrons, especially the backpropagation, are pointed out. Many papers deal with the comparisons between the classical neural networks performances and those of the new ones.

Valentin Drăgoi

DENKER S. J. (Ed.), *Neural Networks for Computing.* AIP Conf. Proc. *151,* American Institute of Physics, New York, Third Printing, 1989, 445 pp., ISBN 0-88318-351-X.

This volume contains the proceedings of the conference on „Neural Networks for Computing" held in Snowbird, Utah, on April 13-16, 1986. There are presented over 60 worthy papers.

The presented volume offers a wide spectrum of neural approaches into the information processing domain. Thus, one could remark many interesting developments concerning higher order corelation models, stochastic models for the pattern recognition problem, applications in digital signal processing, connectionist networks, new learning rules, VLSI architectures for implementation of neural networks, etc.

The reader will be able to find papers which emphasize the use of the artificial neural networks as an intelligent way which try to reduce the complexity of the information processing field. There are presented some extensions of the currently popular models in agreement with the conference topics.

Some of the most important contributors in the neural networks domain such as J. Anderson, G. Carpenter, K. Fukushima, S. Grossberg, G. Hinton, B. Kosko, T. Sejnovski, etc., have attended the conference meetings and offered again a measure of the growth in the field by the level of their recent works.

The proceedings have conjoined the best ideas of the last years thus characterizing in the same time the improvement of the development of real applications such as pattern recognition, target detection, expert systems, learning motor tasks, surface interpolation, modelling cortical functions, etc.

Valentin Drăgoi

KÖSSLING H., PREISSLER H., DONAU M., *Lehrbuch Fernsprechtechnik.* Brandenburgisches Verlagshaus, 1990, ISBN, 3-327-00865-5.

In einem Zeitraum wo wir einer wahren Explosion der Entwicklung der Fernsprechtechnik beistehen, muss die Wichtigkeit dieser nicht mehr betonnt werden. Offenbar, auch die Interesse für Werke die sich damit befassen.

Die Verfassern beabsischtigten in einem einzigen Band möglichst viele Informationen darzustellen. Betrachtet werden also die Grundlagen der Fernsprechtech-Bauelemente und Baugruppen, Fernsprechendtechnik, Trägerfrequenz-Übertragungstechnik, Digitale Signalübertragung, Fernsprechvermittlungstechnik.

Die Absicht der Autoren den Lesern einen geraumen Gesichtskreis zu verschaffen wurde damit gänzlich erfüllt. Die einfache und klare Präsentierung, die gut gewählte Reihenfolge der Kappiteln, das pedagogische Gefühl den Kern der Problemen den Lesern beizubringen machen aus diesem Buch ein Erfolg für die Verfassern, eine sehr nützliche Quelle für eine gewisse Kategorie des Publikums fürs das Buch geschrieben wurde.

M. Antonescu

CLASSEN L., WIESNER U., *Wiessensspeicher 80286 — Programmierung.* VEB Verlag Technik, Berlin, 1990, ISBN, 3-341-00867-5.

Seit mehreren Jahren entwickelt sich die Mikroinformatik in einem beeindruckenden Rhytmus. Zahlreiche Menschen verschiedener Berufen interessieren sich jetzt für die Anwendungen der PC-s. Der Grund dieses Booms basiert sich auf die Existenz starker und kompakter Rechenanlagen, darunter auch die Maschinen die als Zentraleinheit einen 80286 Mikroprozessor haben. Unter diesen Bedingungen ist also wichtig entsprechende Literatur zur Hand zu haben.

Einer logischen Gedankenreihe folgend, beginnt das Buch mit allgemeinen Grundinformationen : Systemübersicht, Betriebsarten, Registersatz, Adressraum und Speicherverwaltung. Die Darstellung setzt fort mit Detailen die sich auf Zugriffschutz im PVA Mode, Taskkonzept PVA Mode, Interrutpsystem ein Ausgabe, Datenformate und Adressierungsarten, Systeminitialisierung und Kompatibilität zum 8086 und 80836.

Die wichtigsten Merkmale des Werkes — darunter die reiche Information, die Klarheit der Darstellung, die gut gewählte Ordnung der Präsentierung — sichern die Interesse und die Schätzung der Lesern.

M. Antonescu

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